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The Proceedings

OF

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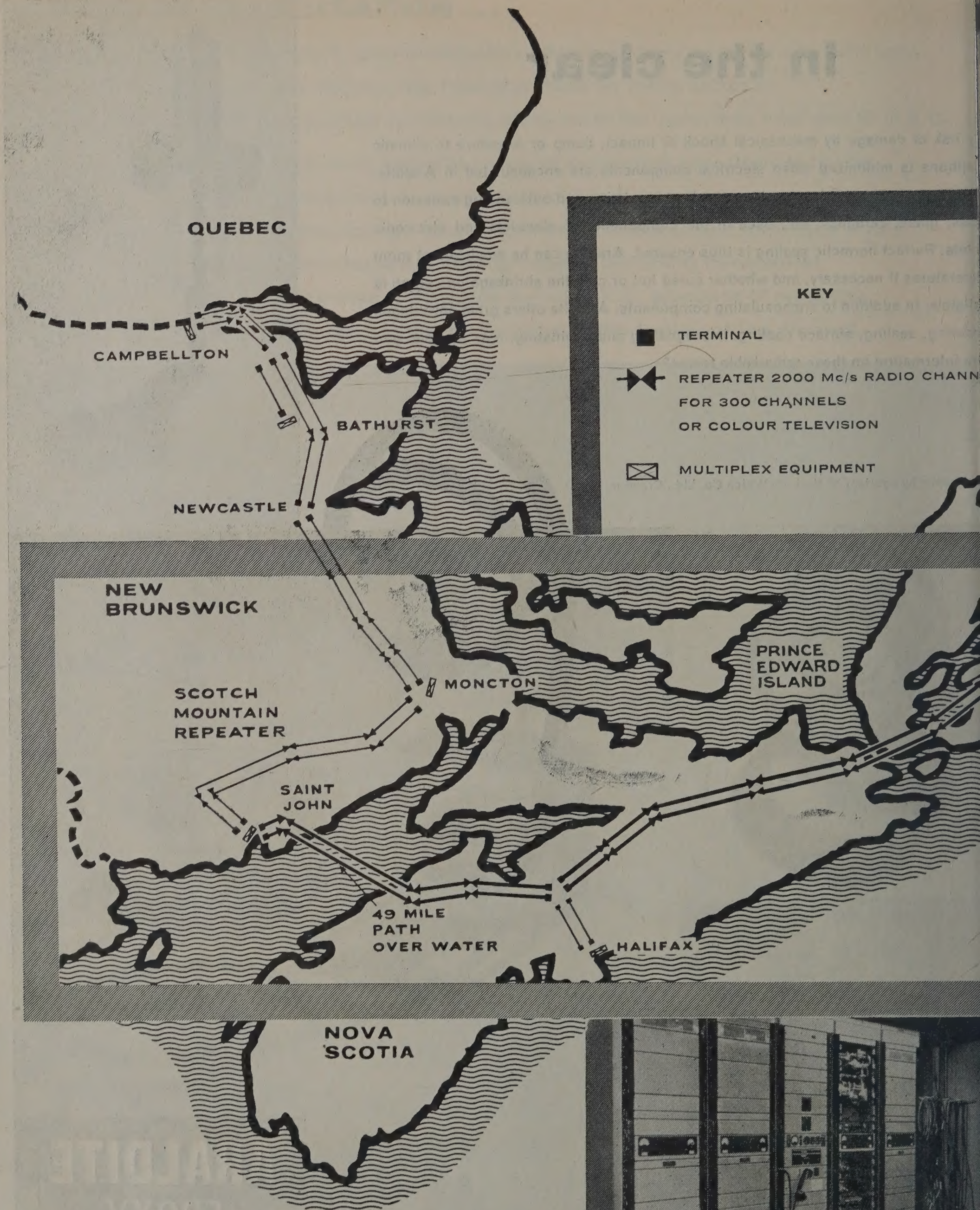

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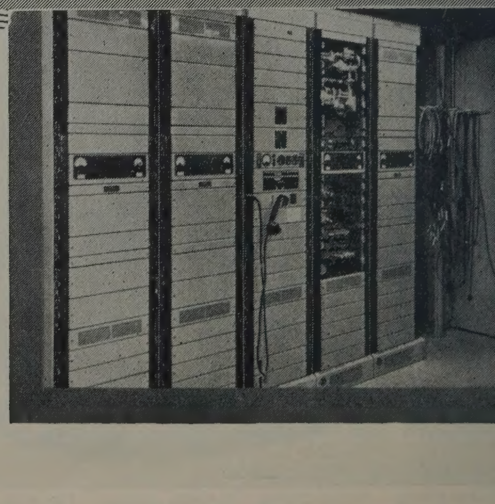

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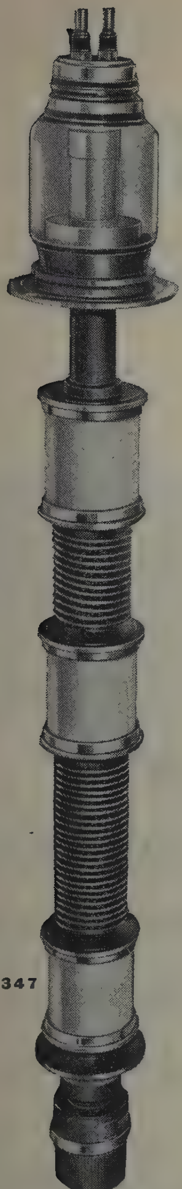

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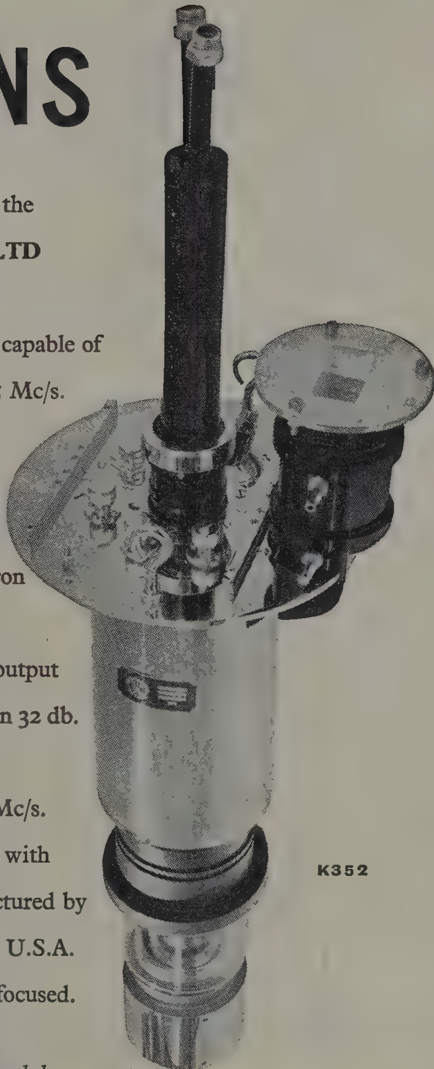

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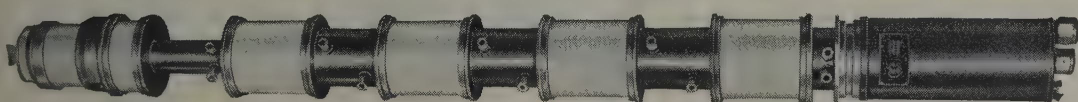

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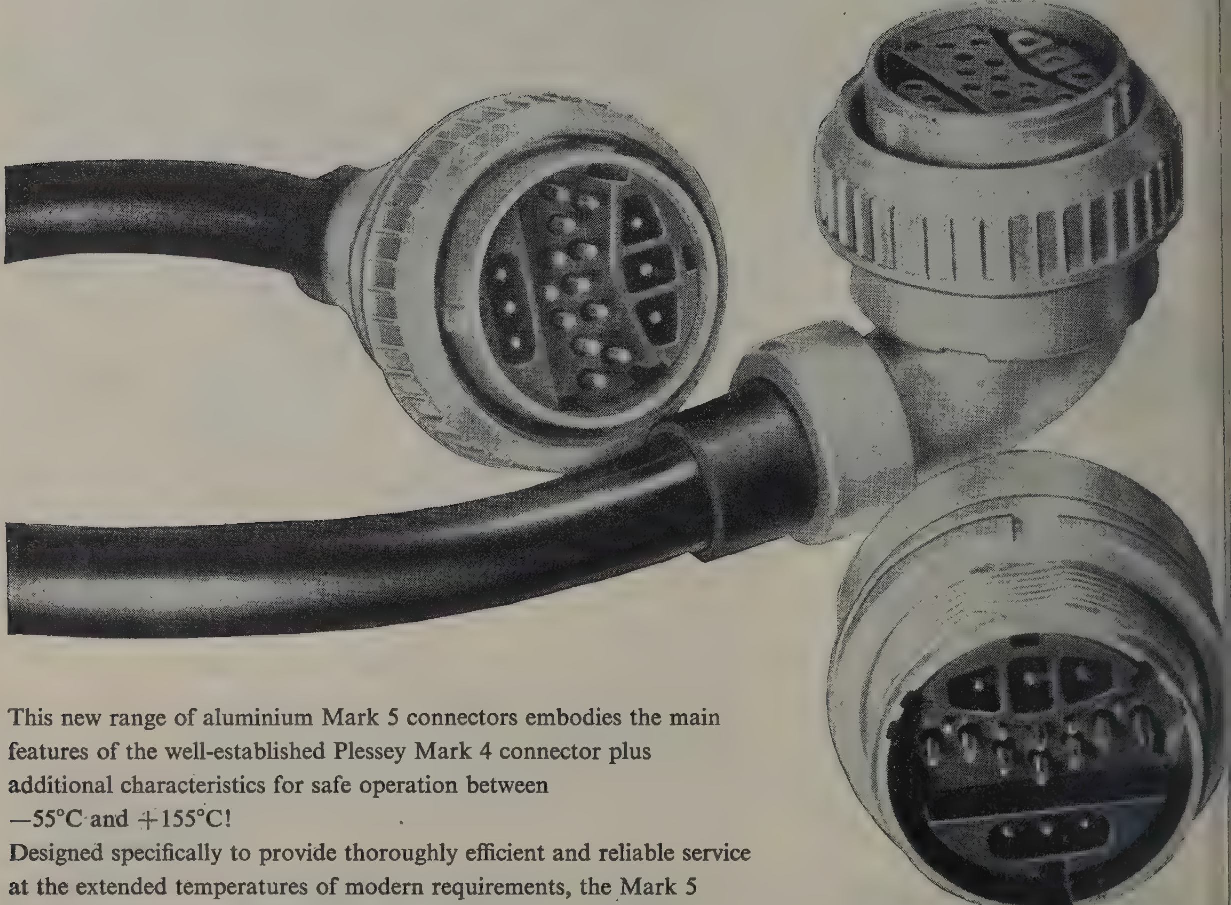

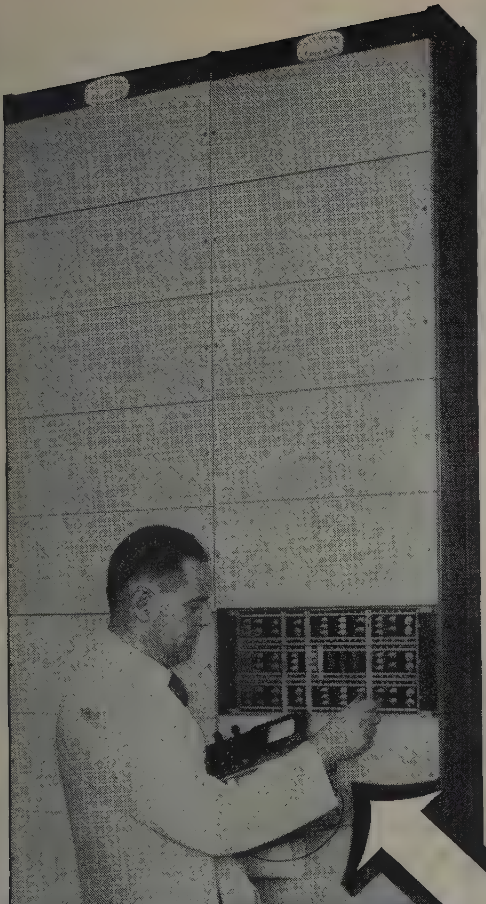

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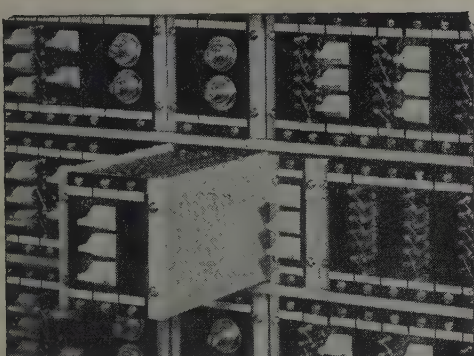

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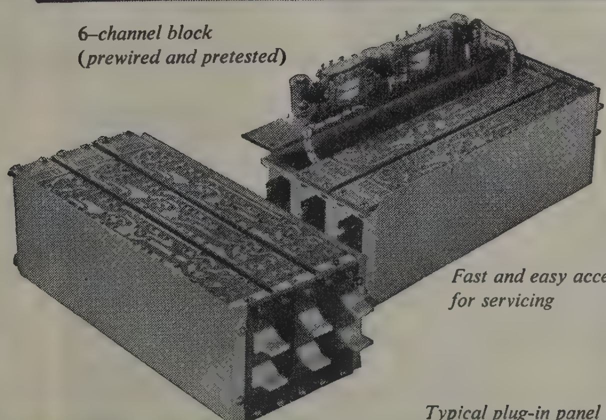

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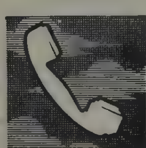
vast experience  **in the science of telecommunicatio**

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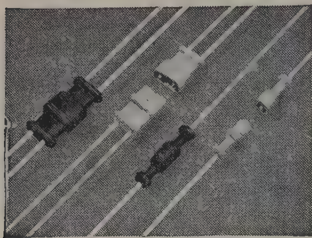

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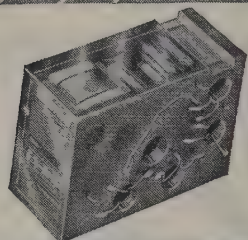

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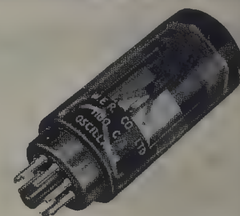

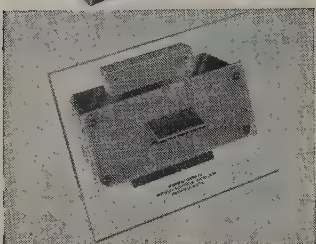

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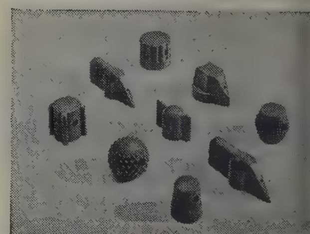

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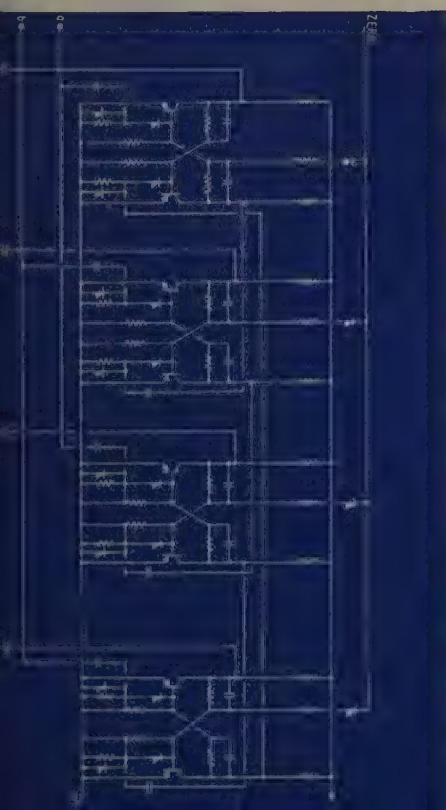
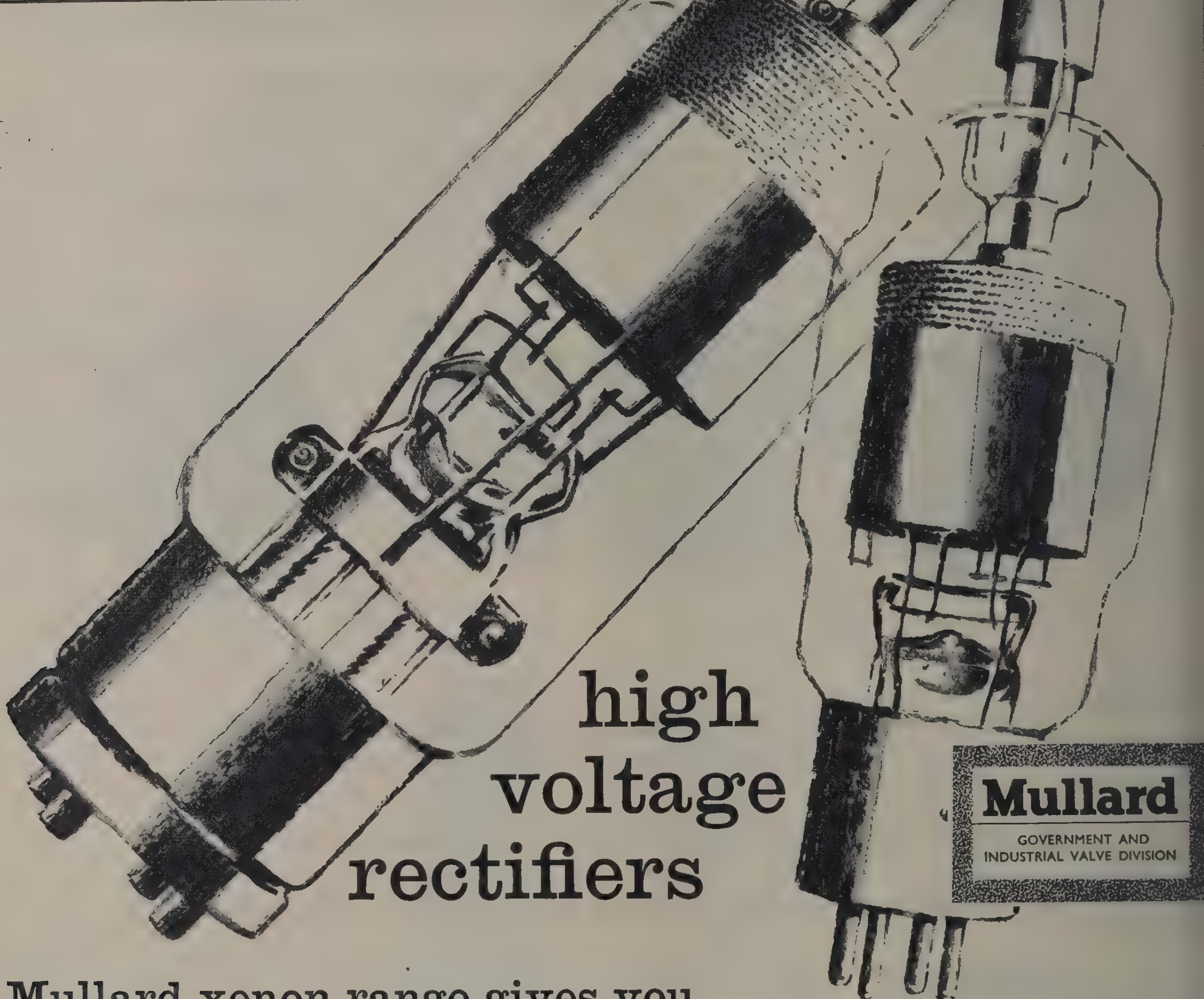


Abridged Data

Type No.	American No.	CV No.	Base	Vf (V)	If (A)	P.I.V. max. (kV)	Ik(pk) max. (A)	Ik(av) max. (A)	Heating-up Time (sec)
RR3-250	3B28	CV1835	4-pin UX	2.5	5.0	{ 10 5.0	1.0 2.0	0.25 0.5 }	10
RR3-1250	4B32	CV2518	B4F	5.0	7.0	10	5.0	1.25	30
RR3-1250A	—	CV2399	Goliath Edison Screw	4.0	11	13	5.0	1.25	30
RR3-1250B	—	—	Goliath Edison Screw	4.0	7.0	13	5.0	1.25	30



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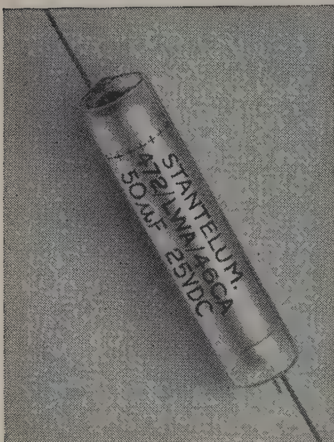

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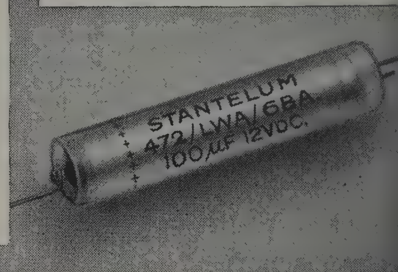

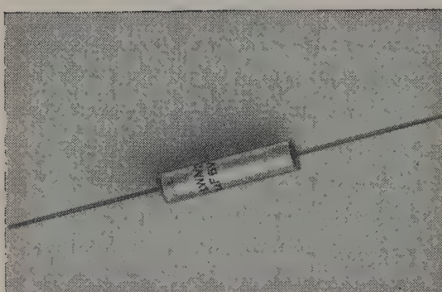

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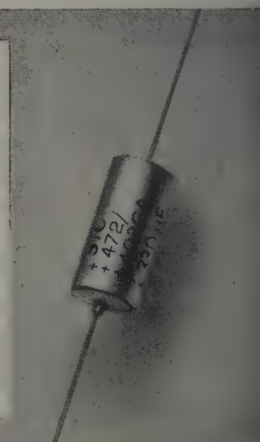

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Paper No. 3213 E
April 1960
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PLANNING AND INSTALLATION OF THE SOUND BROADCASTING HEADQUARTERS FOR THE B.B.C.'S OVERSEAS AND EUROPEAN SERVICES

By F. AXON, D.C.M., Associate, and O. H. BARRON, M.B.E., B.Sc.

he paper was first received 11*th May, and in revised form* 15*th December*, 1959. *It was published in April*, 1960, *and was read before the* ELECTRONICS AND COMMUNICATIONS SECTION 2*nd May*, 1960.)

SUMMARY

llowing a brief history of the B.B.C.'s external services, the paper nes the reasons for the present form of centralization at Bush se. The method of programme building is considered and the io and recording facilities are detailed. The equipment in the rol room and other technical areas is described. Installation lems are discussed and the paper concludes with some notes on tical experience with the installation in service.

(1) HISTORICAL SURVEY

(1.1) Pre-War Services

November, 1927, the British Broadcasting Corporation gurated a series of experimental transmissions of items their programmes on a wavelength of 24 m (12·5 Mc/s) a transmitter of 8–10 kW at Chelmsford. In view of the t interest in these experimental transmissions, especially in Dominions and Colonies, it was decided that a more comensive short-wave service was justified. The original Empire ice, in English, was opened on 19th December, 1932, on transmitters of 10–15 kW installed at Daventry, with a plement of six directional and six omnidirectional aerials.

single programme of some ten hours per day was radiated. ount was taken of the differences in longitude of the various s of the Dominions and Colonies by suitable timing of the smissions.

uring the early years the arrangements for the presentation his service were relatively straightforward, as it ran more or in parallel with the domestic service. A single line feed to entry was adequate; many of the items were radiated 'live' multaneous broadcasts from the domestic services.

ne major departure was the recording of programmes at the of the live broadcast, for reproduction to different areas in transmissions. This overcame the difficulties arising from use of directional beams, which served only a limited target at one time, and also the fact that peak local listening times ed widely in Greenwich mean time. This necessitated a considerable expansion of both recording and reproducing facilities, and much experimental work was undertaken on systems other than disc, e.g. Marconi-Stille, Philips-Miller, etc. Facilities were also provided for additional news bulletins.

With the worsening international situation in 1938, programmes to foreign countries, mainly news bulletins, in languages other than English were introduced. Studio space was provided at Broadcasting House.

(1.2) Wartime Expansion

By the end of 1939, bulletins in several additional languages had been added. Separate Empire and European services were now envisaged and additional studio facilities were obtained by the use of existing studio premises at Maida Vale.

During 1940, further bulletins were introduced, making, in all, programmes in 43 languages.

More studio facilities were needed and during 1939–40 the major part of the Empire service was evacuated to Wood Norton, near Evesham. The Empire and European services continued to expand and ultimately became:

Empire Service.—(*a*) World service in English for 20 hours per day, divided into four main transmission periods.

(*b*) More specialized regional services in English, for some 8½ hours per day, for the Commonwealth countries.

European Service.—(*a*) Mainly Central European languages for some 18 hours per day.

(*b*) Mainly Balkan languages for about 8½ hours per day.

Further studio facilities were provided for the Empire service by developments at Wood Norton, and for the European service by the conversion of offices in Bush House into a small control room with news studios and recording rooms.

In 1940, additional premises were taken over near Wood Norton and equipped to provide studios and a control room for one network. A significant point in the operation of these studios was the close association, encouraged by the juxtaposition of the control room and studios, which developed between the control room and studio staff. This close method of programme operation later developed into the form of con-

The authors are with the British Broadcasting Corporation.

tinuity suite working used at Oxford Street and, for certain networks, at Bush House.

With the increasing pressure on studio requirements and the threat of enemy bombing, plans were drawn up for further expansion by the use of premises at Aldenham for the Empire service (December, 1941), the conversion of a large store at 200, Oxford Street for the Empire service (May, 1942) and the extension of Bush House facilities for the European service (end of 1942).

(1.3) Programme Distribution

This rapid expansion gave rise to many problems concerning the technical operation of studios, lines and transmitters, due to the use of languages other than English. These were solved by introducing 'colour coding' in an elementary form throughout the system during 1939–40.

The short-wave transmitters, situated in different parts of the United Kingdom, were connected to the studio centres by Post Office lines. The number of lines to each transmitter site was related to the number of 'colour networks' radiated simultaneously. Programme duration was standardized on the basis of multiples of one quarter-hour (Section 3.1) and the separate language programmes became contributions to the relevant colour networks. Each line was scheduled to carry one or more colour network(s) in succession, the complete 24-hour sequence being designated a 'chain', to which a number was allocated.

At the sending end, each colour network had to be connected to the appropriate chain at the correct time. At the transmitter sites, each chain had to be connected to the correct sender(s). (The sequence of colours in a chain was not necessarily the same as the sequence of colours required by a particular sender.)

In May, 1943, all programmes for the external services were routed into Aldenham, which became the sending end. The time available for switching the colours at each quarter-hour being only a few seconds, it was necessary to preset all chains and switch them simultaneously.

Two bays of jackfields were provided, designated X and Y, each equipped with jacks for colours and jacks for chains. The required set-up was plugged for the current quarter-hour on one bay and for the following quarter-hour on the other. Either the X or Y bay was 'on transmission', the chains being connected to the lines by relays. At the appropriate time, the operation of a key changed the lines from one bay to the other. The engineer then had a quarter-hour in which to replug the bay which had come 'off transmission'. Comprehensive monitoring was provided.

The switching at the transmitters was done manually at each quarter-hour, if required. This manual switching has since been replaced by automatic switching at some transmitters.

(1.4) Post-War Concentration

At the end of the war, all overseas services were concentrated at Bush House and Oxford Street, and by May, 1952, the studios at Aldenham and Wood Norton had closed down, the switching bays having already been transferred to Bush House. Programmes originating in Oxford Street were sent to Bush House for distribution to the various transmitters.

Generally speaking, this was the position until the new control room at Bush House was commissioned in November, 1957.

(2) CONCENTRATION AT BUSH HOUSE

(2.1) Alternative Proposals for New Control Facilities

The proposed concentration of the external services at Bush House was first considered in 1949, in anticipation of the eventual de-requisitioning of 200, Oxford Street which finally occurred in December, 1957.

Three alternative methods of expanding the control-r facilities at Bush House were discussed:

(*a*) To enlarge the existing control room in the basement ar
(*b*) To provide a new control room to handle the Overseas Se transferred from Oxford Street, leaving the old control roo carry European services, as formerly.
(*c*) To provide a comprehensive control room capable of ac modating all services.

The outstanding arguments in favour of a comprehe control room, leading to the adoption of (*c*), were:

(i) It was estimated that there would be a saving in sta 25 compared with either of the other proposals.
(ii) Although the estimated capital cost was substantially h than the alternatives, it appeared that revenue savings would this in less than three years.
(iii) On completion, the equipment would be up to date in of comprising obsolete or hybrid designs.
(iv) A control room above ground level would be an impo factor in improving working conditions for staff, apart from stimulating effect of handling equipment of modern design.
(v) There was considerable doubt as to the availability of suff space to expand the old control room.

Accordingly, in June, 1954, a specification for a new co room and ancillary areas in the centre block was prod Design work and planning was started on the studios, reco areas and continuity suites, in addition to the control itself. The new studios were to be accommodated in the c block.

(2.2) Method of Switching*

Three basic systems were considered: plugs and jacks, r and motor uniselectors. Plugs and jacks were in use i existing control rooms at Oxford Street and Bush House. provide complete flexibility at a low capital cost but are u and uneconomical in operating staff for a large install Relays are satisfactory for a small or medium-sized control and are used by the B.B.C. under these conditions. F installation of the size of Bush House, the number of would have been prohibitive.

Motor uniselectors provide a more economical system relays for large installations and were the choice in this inst They provide a ready means for simultaneously switchin programme circuit and such ancillaries as the control lin programme, etc. Sixteen-level switches are used, each 50 inlets.

Source switches connect a selected source to a channel th the switching system; route switches connect a selected ch to a destination.

A switch is operated by 'marking' an inlet with 'batte one level and operating a key to motor the switch. Cont cubicles (Section 3.2) have access to a source switch for programme channel and individual marking for these swi A miscellaneous switching position has access to a large nu of switches and uses common marking. This means th the switches are marked simultaneously, but only the re switch is motored by operating the appropriate key.

(3) METHOD OF WORKING

(3.1) Nature of Programme Building

Basic Programme Structure.—For reasons of operationa venience, it has long been the practice in External Servi standardize programmes in multiples of a quarter-hour. programme starts at a 'zero', corresponding with an quarter-hour G.M.T., and finishes 20 seconds before a s quent quarter-hour zero. After switching operations have

* For full details see PETRIE, R. D., and TAYLOR, J. C.: 'Programme Sw Control and Monitoring in Sound Broadcasting', B.B.C. Engineering I Monograph No. 28, February, 1960.

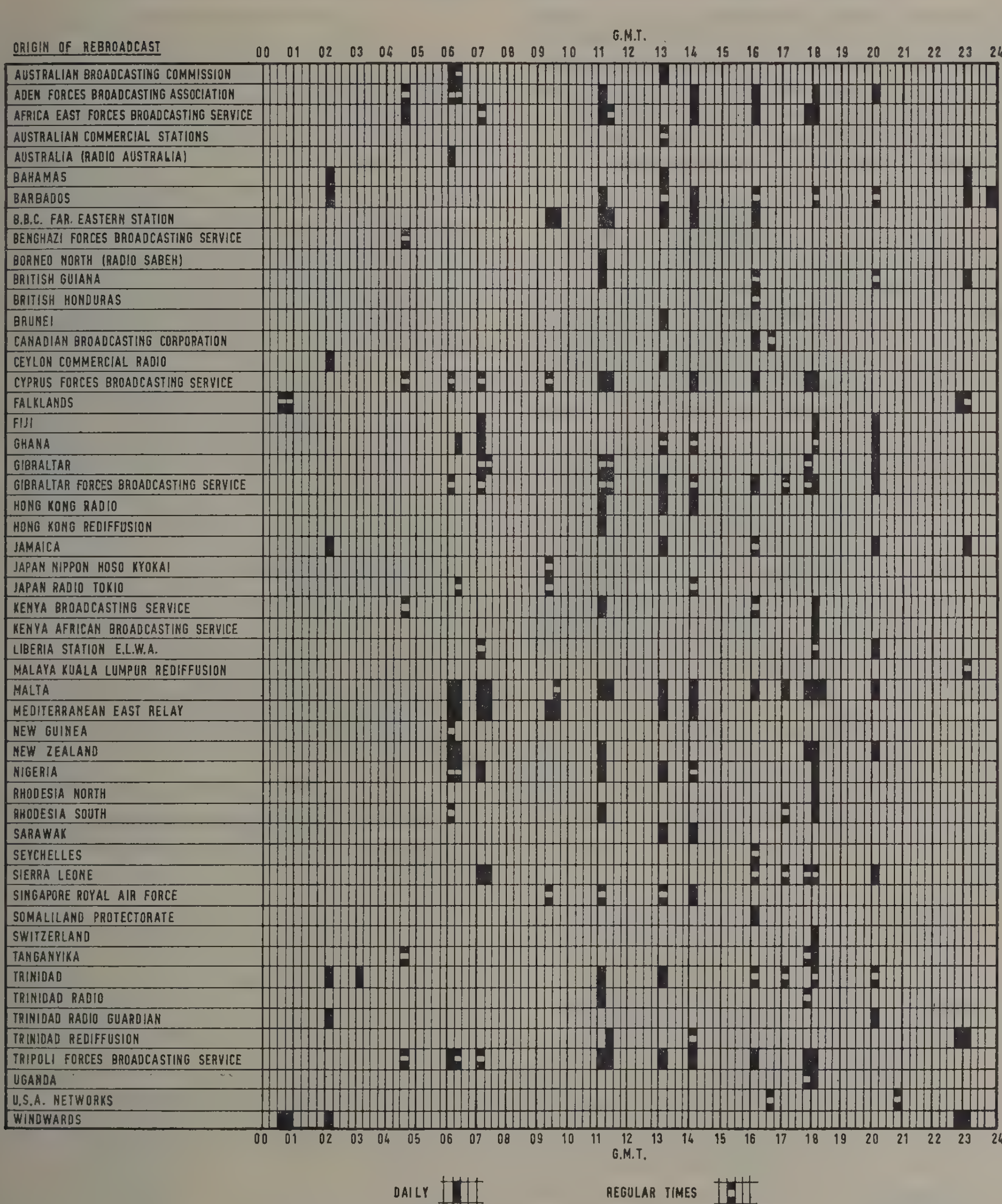

Fig. 1.—Regular rebroadcasting of General Overseas and North American Services news periods throughout the world, May, 1958.

pleted, the remaining period prior to zero may be used for signals or network interval signals.

lation of Programmes and Networks.—The allocation of age programmes to colour networks depends mainly on rs of engineering convenience, such as the manning of the nuity positions at the time and the availability of certain os. The same language may appear in various colour orks at different times of the day, but it is common to find ages associated with a particular geographical area running after another in the same network.

ming.—The importance of strict timing is obvious when global scale of the complete operations is realized. B.B.C. rammes are likely to be rebroadcast and the overseas organizations concerned must be able to rely on exact timing. Some idea of the extent of such rebroadcasting can be obtained from Fig. 1 which shows regular rebroadcasting of the General Overseas (G.O.S.) and North American Services (N.A.S.) news bulletins during May, 1958.

Non-Standard Programme Periods.—There are a limited number of variations of the quarter-hour principle. The most important of these are the Programme Parades which are designed to sign-post the G.O.S. transmissions. The need for this arises out of the almost continuous nature of the service, during which coverage changes from area to area are necessary in accordance with local listening times.

Programme Parades are mounted as separate programmes

prior to the start of G.O.S. transmissions directed to new audiences. They give the listeners up-to-the-minute details of subsequent programmes, as well as any changes in the published wavelengths, before the transmissions concerned join the main programme. Their duration varies from 10-minute introductions when the service re-opens after a close-down period, to a mere 1½ minutes prior to the start of new area coverage.

(3.2) Continuity Working

Continuity working, which is used throughout the B.B.C.'s sound services, takes two different forms in the External Services. Where a network is continuously broadcast to the same group or groups of countries, the programme material will come from a variety of sources and a continuity suite, comprising a studio and cubicle, is employed. All sources are fed into the cubicle, together with linking announcements from the associated studio and fill-up records, if necessary, to form a continuous programme.

Where, however, a network consists of a number of short programmes, each directed to a different part of the world, no continuity is involved and a continuity studio would be superfluous. In this case the continuity cubicle 'assembles' the network from the individual items, with no linking announcements.

In each case, the complete network is fed from the cubicle to the control room and thence to the chain switching equipment.

A special feature of the G.O.S. is that all announcements, together with the last few seconds of the preceding and the first few seconds of the following programmes, are recorded automatically on tape.

(3.3) Chain Switching; Automatic Switching Unit (A.S.U.)

Although manual plugging at some transmitters was replaced by automatic switching (Section 1.3), the original chain switching equipment was still in use at Bush House. An obvious development was to use automatic switching for this purpose also. (Section 5.5.)

(3.4) Miscellaneous Requirements

The main technical functions of Bush House are to asser the complete networks in the various continuity positions an distribute these networks as chains to the various transmit by means of the a.s.u. There are a number of miscellan requirements that account for a large proportion of the wor the control room. In addition to providing feeds to the sl wave senders, there are several other destinations:

(*a*) Continental Trunk Exchange, for programmes to fo countries, frequently as part of a two-way programme.
(*b*) Recording suites. (Sections 4.2 and 7.)
(*c*) Offices, to enable programme compilers to check their grammes and to listen to others which have a bearing on work. This is known as the 'programme ring-main'.
(*d*) A selector system, which enables the B.B.C. receiving st at Tatsfield to dial any network and then compare the ori via a direct line from Bush House, with the radiated programm

News despatches are received from abroad and these ma routed direct to recording channels, for eventual transmi or to office-type recording equipment, for subsequent ty and later use in news bulletins or other programmes.

Items are picked up by the Tatsfield receiving station an mixed with other material in studios or continuity position re-radiation.

Certain studios have facilities for adding echo to progran There is one echo room and also an artificial-echo mac The studios concerned can select either source of echo appropriate circuits guard against two or more studios connected to an echo source at the same time.

A loudspeaking intercommunication system is pro between the technical operations manager, who has a c in the control room, and the various continuity cubicles.

(3.5) Block Schematic (Programme)

The present installation is for a maximum of 150 source provision has been made for expansion to 200 sources. are 72 channels through the switching system. Since cont

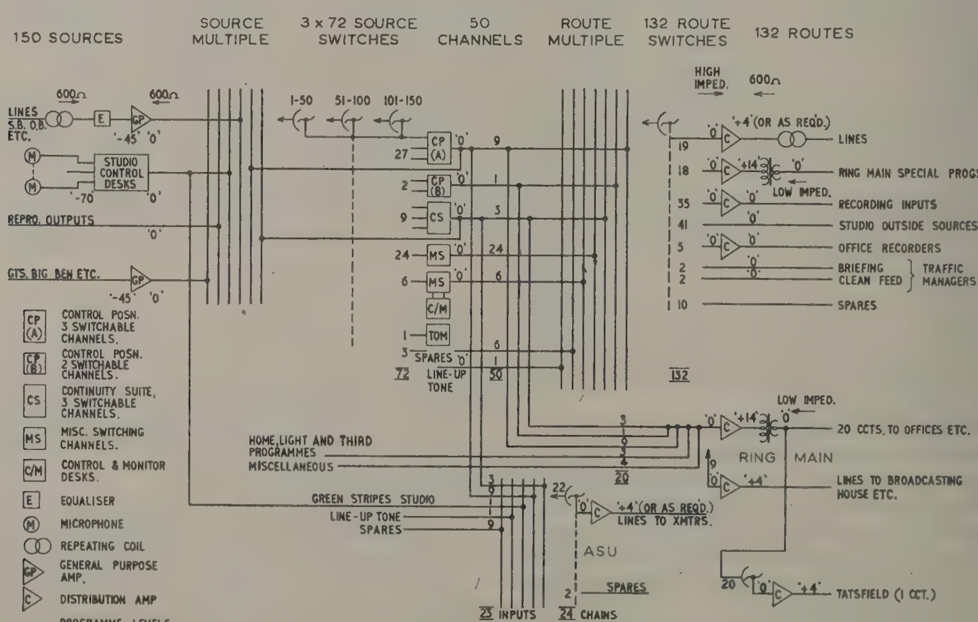

Fig. 2.—Programme switching block schematic.

tions have more than one input channel but only one outlet, e are less than 72 channels in the route multiple. Forty-four n use and these are accommodated on a single route switch each destination. In addition to feeding into the route iple for switching, the outputs of continuity positions are ired for a number of permanent connections. The a.s.u., ring main and the Tatsfield selector are wired across the opriate outputs.

g. 2 is a block schematic of the programme switching system. lst much of this will be readily understood from the fore- g description, a few points require further explanation.

f the 30 miscellaneous switching channels, 24 are, in effect, ole-ended cords between a source and a destination. The aining six are wired to control and monitor positions.

ne technical operations manager is responsible for all tech- l equipment in Bush House, and has access to the source iple for monitoring purposes. He also has access (not vn in Fig. 2) to the outputs of continuity positions.

ach of two traffic managers, who deal with news items and atches, uses two destinations from the route multiple. The fing circuit' will be switched to a correspondent at some ote point and will also be connected to the traffic managers ews rooms in other B.B.C. premises. The 'clean feed' will witched to the remote correspondent only and can be used arry news items for recording. The traffic manager has ote-start facilities for a tape recorder in one of the recording nnels.

THE TECHNICAL ARRANGEMENTS AT BUSH HOUSE

(4.1) Studios

xcluding the continuity suites, there are 31 studios, including new ones in the centre block. These are equipped with the st designs of studio apparatus (Section 6.1). For reasons of nomy, several of the original studios retained their old equip- ts, suitably modified to take advantage of the new control m facilities (Sections 6.2 and 6.3). In the centre block offices are arranged around the outside of the building, iving natural light and ventilation. Studio suites are accom- lated on either side of the central staircase and lifts. Each lio, with its associated cubicle, leads off a vestibule, the two rs thus provided forming a sound trap when staff move veen studio and cubicle. The normal equipment comprises pe B control desk (Section 6.1), three 78 r.p.m. twin-turntable s of B.B.C. design and two reproducers for 45/33⅓ r.p.m. -groove records.

he control room and switching areas, constructed on the n floor of the centre block, provided a problem because of heavy loading imposed on the floors by the equipment and le runs. It was necessary to construct a reinforced-concrete over the existing floor slabs to transmit the loading on to steel beams and stanchions, which were strong enough to port the additional weight. The ceiling of the control room covered with perforated fibre-board tiles to reduce the e level.

he new studios, accommodated on five floors, are built in middle of each floor between the internal stanchions with a ulating corridor to give access to the studio suites and the es. They are mechanically ventilated by a plenum system, air being washed and tempered.

laborate precautions were taken to ensure adequate sound lation between the studio suites and the rest of the building. er exhaustive tests on the existing structure by the B.B.C.'s earch Department, it was decided to provide 'floating' floors hese areas, constructed in light-weight concrete resting on a tress of glass fibre.

The wall construction was designed as two separate skins; either 4½ in brick, breeze concrete or stud partitioning faced with building board was used, depending upon the sound insulation required. The inner skin was isolated from the floor by 1 in cork, and the outer skin, in addition, was isolated from the columns which it abutted by ⅝ in building-board strips. The studios were provided with plasterboard false ceilings. This construction, which is of a light-weight nature, was partly dictated by the limited permissible floor loading, but it achieved the necessary sound insulation. The observation windows were formed in two thicknesses of plate glass, separated by an air space of approximately 8 in. Steps were taken to prevent nearby impact noises from being transmitted to the studios by providing a sponge-rubber underlay to the floor in the corridor which surrounds the studio areas.

Most of the studios and control cubicles are similar in size (approximately 3 000 ft^3) and acoustic treatment was designed to achieve a flat reverberation time of 0·35 sec. Two types of absorbers were employed, the first being of a porous nature (rock wool), and the second being membrane resonators (a type of absorber for bass frequency evolved by the B.B.C. Research Department). The porous absorbers function as a result of the friction produced by movement of air particles through tightly packed fibrous material 1 in thick; they absorb well at frequencies where the depth of material is significant in comparison with the wavelength, i.e. 500 c/s and upwards. The membrane unit acts as a mass attached to a damped spring and achieves maximum efficiency at a frequency determined by the mass and the stiffness of the spring. It is constructed by using roofing felt as the membrane (this supplies the mass and most of the damping) over an enclosed air space which acts as the spring. The resonant frequency can be varied by altering the volume of the air space. Reverberation measurements were taken on the completion of the acoustic work and some adjustment was made by adding shallow roofing-felt units to increase the absorption between 250 and 700 c/s. The acoustic condition of all the studios was then found to be satisfactory, although most of them had more absorption at 1 000 c/s than had been expected. It was thought that this was due partly to the exceptionally efficient absorption by carpets at this frequency and partly to the fact that most of the perforated decorative covers to the acoustic material produced a rather weak Helmholtz effect, resonating at about 1 000 c/s. One result of the Corporation's experience in these studios is that the Research Department has since designed an 'anti-carpet' absorber which absorbs very well at frequencies below and above 1 000 c/s, but it is not so efficient at that frequency.

For covering the acoustic absorbers in the studios, the minimum amount of perforated and slotted hardboard was to be used, as these materials are aesthetically unacceptable. Consequently, curtain fabric, either draped or stretched in panels, was used in the studios instead of slotted hardboard. It was found convenient to use the fabric colour as the dominant colour around which the decorative schemes were devised for each studio suite.

(4.2) Recording/Reproducing Suites

Extensive tape recording and reproducing facilities are available, including two central tape rooms with facilities for remotely controlled recording or reproduction of 20 programmes simultaneously. For administrative and operational convenience, one room is generally used for recording and the other for reproduction, but each room may be used for either or both simultaneously.

Seven separate tape channels, three of them also equipped with disc recorders, are available for use as recording, editing or dubbing channels; three of these channels are provided with

facilities for reproducing or recording at speeds other than standard in order to deal with tapes which may have been recorded in the field at incorrect speeds. In addition, five of the separate channels are provided with remote-control facilities for reproducing and are occasionally used for reproduction of a programme requiring individual attention which cannot, of course, be provided in a central tape room.

There are three disc-recording and one disc-reproducing channels; these, in conjunction with disc-reproducing desks in studio cubicles and disc recorders in three of the editing rooms, provide adequate facilities for this method of recording, the need for which has undergone a drastic contraction with the advent of tape.

Two rooms have been equipped, one as a quality checking room, with tape and disc recorders and reproducers and facilities for the correction of frequency response, and the other as a special-projects recording room, also with tape and disc recorders and reproducers. This latter room, in addition to the normal facilities, is provided with tie lines to, and remote control from, two adjacent studio cubicles and is used for complicated programmes involving material recorded during the programme itself.

(4.3) Continuity Suites and Cubicles

Three continuity suites have been provided for the 'continuous programme' networks—Green (General Overseas), Purple (N. America) and Brown (Arabic). A studio between the Green and Purple suites, designated 'Green Stripes Studio', is used for opening, closing and special announcements to countries joining or leaving the Green network.

For the networks comprising a number of independent programmes, ten continuity cubicles have been built adjacent to the control room.

(5) CONTROL-ROOM SUITE

(5.1) General

The sixth-floor control-room suite comprises a numb
separate areas (Fig. 3).

The control room, of approximately 1 700 ft^2, contain
operating desks and bays of equipment.

The auto-switching equipment and the main distrib
frame are housed in a separate room, not normally st
Precautions are taken to keep this as clean as possible.

The a.s.u. equipment is divided between this room and a
cubicle, partitioned off from it.

The clock equipment, a 50-volt battery and the asso
charging equipment are housed in small areas adjacent t
switching room.

(5.2) Desks and Auxiliaries

The main operating position consists of a large desk wi
main indicator over the centre and a supplementary ind
over the left-hand end.

Three double 'control and monitor' positions each cons
a separate desk. There is also a cubicle for the tec
operations manager, with a good view of all operation posi

Miscellaneous Switching Position.—This is at the left-han
of the main operating position and is used for setting u
large number of miscellaneous circuits. It consists of 'ma
keys for sources and destinations and 'operate' keys f
channels. Keys associated with the supplementary ind
complete the equipment.

Main Indicator.—This is a large display over the entire
part of the main operating position and shows all the sou
channel—destination connections which are set up. It co

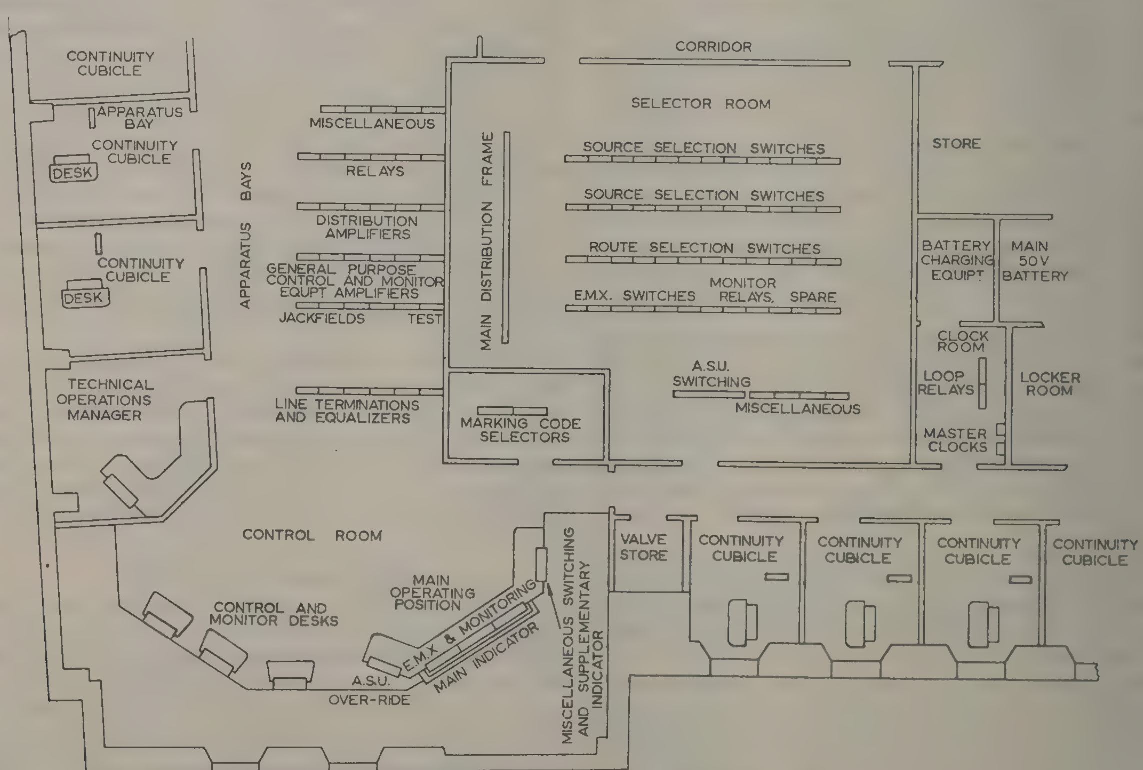

Fig. 3.—Layout of control room and associated technical areas.

·› sets of windows, the first comprising one window for ·hannel and the second, one window for each destination. · a source is selected to a channel, the source code lights in ·propriate channel window. When the channel is selected ·stination, the channel number lights in the corresponding ·ation window.

·plementary Indicator.—Although the main indicator gives ·plete picture of all the connections set up, a considerable ·t of searching is required over a fairly large area to ·ine the source and destination connected to a particular ·el. The supplementary indicator, which is a small display ·the miscellaneous switching position, overcomes this ·lty. A key on the desk is associated with each channel ·ere are three sets of windows in the indicator, i.e. sources, ·els and destinations. Operation of a channel key lights ·rresponding channel window, the source connected to it ·y) and the destinations (if any). It also provides aural ·oring on the source.

display is cancelled when another channel key is operated. only one display can appear at a time, in contrast to the ·splay of the main indicator.

·ineering Manual Exchange (E.M.X.).—A 200-line cordless occupies part of the centre portion of the main operating ·on, 12 tie circuits being provided. All subscribers have · to one another, and, when a connection is made, the codes two participants are displayed in a window associated with ·rticular tie circuit in use.

·nitoring Positions.—Visual and aural monitoring can be ·d out at key points on all circuits. Two identical moni· positions are available, one on each side of the e.m.x. ·nables one operator to look after the e.m.x. and monitoring programme activity is low, but, in peak periods, two can the e.m.x. and each has full monitoring facilities.

·omatic Switching Unit Over-ride.—Provision is made to ·ide the a.s.u. to cover such eventualities as last-minute ·es in the programme schedule, fault conditions, etc. · and aural monitoring of chains is also provided.

·scellaneous Position.—Miscellaneous equipment comprises:

A set of red lights to indicate which studios and reproducing ·els are on transmission.

Keys, microphone and loudspeakers for two separate ·peaking intercommunication systems to continuity cubicles few miscellaneous positions.

An alarm panel with lamp indication of various fault ·tions.

·ntrol and Monitor Positions.—Each control and monitor ·on provides two entirely separate and identical operating ·ons. The appropriate channels from the miscellaneous ·ing position pass straight through these positions until they ·rought into circuit, by the operation of keys. Normal ·l facilities are then available, including a channel fader, and aural monitoring, pre-fade monitoring, control-line ·ies, remote start of a recording and transmission of line-up ·nd interval signal.

·hnical Operations Manager's Desk.—The technical opera· manager requires comprehensive monitoring and com·cation facilities. For monitoring, he is provided with four ·programme meters, three of which can be connected to the ·ts of any three continuity cubicles. The fourth meter can ·ected to any source. His loudpseaker can be connected · one of the four programmes. For communications, he ·onnect his telephone across the telephone line from a ·d source. He also has two direct lines to e.m.x. In ·on, he can listen across or speak to any subscriber on the ·oudspeaking inter-communication systems. He has a ·ate set of studio red lights.

(5.3) Control-Room Bays

Switching bays should be installed in a dust-free area to avoid contact troubles. Also, on account of their noisy operation, it is desirable to segregate them from the control room. It was therefore decided to have a separate area for switching and other non-operational bays and to include, in the control room itself, only operational bays. Plinths were laid down for 35 bays.

Line-Termination Bays.—Post Office cables are extended via a small distribution frame to three line-termination bays, consisting of U-links and associated repeating coils for incoming music, outgoing music and control circuits. By suitable jumpering, the individual pairs are brought up in the required positions on the line-termination bays.

Equalizer Bays.—Permanent incoming music circuits (for simultaneous broadcast working) are equipped with fixed equalizer/attenuators to give a standard level of −45 dBm at the equalizer output. The fixed equalizer bay consists of 'boxes' into which the equalizers slide, being connected into circuit by means of U-links, to facilitate changing an equalizer.

Occasional circuits for outside broadcast working are continually being changed and are used only a few at a time; variable equalizers are plugged up as required. The outside broadcast equalizer bay is equipped with four different types of variable equalizers which, used in various combinations, are matched to the occasional circuits.

Amplifier Bays.—All the amplifiers are small units, of similar external appearance, plugged into general-purpose mountings, each accommodating ten working amplifiers plus one spare and a rotary-switch unit (manually operated) to insert the spare where it may be required. Two of these mountings are accommodated on each side of the bay and a separate mains unit is used for each.

There are 88 general-purpose amplifiers for level raising and 220 'C' amplifiers for programme distribution. Three bays are equipped with amplifiers and associated relays for the control and monitor positions.

Relay Bays.—The control desks for all studios provide facilities for mixing outside sources with the studio microphones. When an outside source is selected, the control line is extended to the desk and is also used to feed back cue programme, the necessary switching being carried out by relays in the control room.

A standard relay mounting is used which will accommodate up to 25 relays. For the large number of outside source lines, six bays of relays are required, each with ten relay mountings.

Test Equipment.—Comprehensive a.c. test equipment is provided on a single test bay. It comprises a variable-frequency tone source, a fixed-frequency tone source (for line-up purposes), a high-input-impedance amplifier-detector (for level measurements), a test-programme meter and weighting network (for programme or noise measurements), a variable attenuator and a high-pass filter (for measurements of total harmonic content). D.C. testing of lines is carried out with a bridge Megger tester.

Jackfield Bays.—Although the equipment has been designed so that normal switching is carried out by remotely operated motor uniselectors, facilities must exist for over-plugging for special purposes or under fault conditions. Listen jacks are provided at numerous points in the circuits. The number of connections taken through the inner springs of jacks in programme circuits is kept to a minimum.

Programme Ring-Main Distribution Bay.—The purpose of this jackfield bay is to segregate the circuits to different areas, so that a fault at one location does not affect the whole system, and to enable special programmes to be plugged to individual lines. [Section 3.4(*c*).]

Miscellaneous Bays.—Miscellaneous equipment is mounted on five bays in the back row, including echo-switching relays (Section 3.4), continuity tape recorders (Section 3.2), an endless-loop tape reproducer for the 'V' signal used during intervals and aerial amplifiers for a number of television and v.h.f. sound receivers in various offices and other areas.

(5.4) Switching Equipment

Plinths were laid down for 48 bays.

Source-Selection Bays.—A bay of 12 switches makes 50 sources available to 12 channels. A group of three such bays is required for 150 sources and six groups provide for 72 channels. Adjoining each group of three bays, a space has been left for a fourth bay for possible expansion to 200 sources. Associated relays are mounted on the same bays.

Route-Selection Bays.—As described in Section 3.5, each destination requires a single switch, making 11 bays necessary for 132 destinations. Space has been left for expansion.

E.M.X. Switching and Relay Bays.—E.M.X. circuits do not require 16-level switches, but by employing single-ended wipers, eight of which are 180° out of phase with the remaining eight, 100 subscribers can be accommodated on one switch. The contact banks are covered twice, once by each set of eight wipers. The Bush House e.m.x. is designed for 200 subscribers with 12 tie circuits, the switches being accommodated on four bays. There are three bays of associated relays.

Relay Bays.—The relays associated with the comprehensive monitoring system are accommodated on two bays, similar to the relay bays in the control room.

Miscellaneous Equipment.—This comprises the equipment to enable Tatsfield to select a ring-main programme [Section 3.4(*d*)], the a.s.u. (Section 5.5), a Post Office line-termination distribution frame, the main distribution frame on which all the main jumpering is carried out, and the technical operations manager's bay, mounting all amplifiers and relays used with his desk.

(5.5) Automatic Switching Unit

The a.s.u. enables a 24-hour pattern of colours-to-chains to be set up in advance and the appropriate switching takes place at every quarter-hour, initiated by the master clock. It comprises two entirely separate sets of motor uniselector switches—marking code selectors and destination switches.

The marking code selectors advance one position every quarter-hour. A single switch covers 24 hours in the same way that an e.m.x. switch covers 100 subscribers. Each 'marks' a corresponding destination switch in the same way as a switch is marked manually from a continuity cubicle.

'Combs', consisting of insulated mouldings carrying springy-metal contacts, are inserted between the inlet tags of a marking code selector. Each colour programme has a comb with a different arrangement of contacts, these being connected together within the comb. Thus, for each colour, a different combination of levels in the selector is strapped, and this is translated by relays so that the appropriate inlet to the destination switch is marked.

Depending upon the arrangement of combs in the selectors, the destination switches will motor to the correct positions when the necessary operating pulses are applied under the control of the master clock. There are 24 selectors controlling the destination switches for 24 chains.

The bays of selector switches are installed in a cubicle partitioned off from the main switch room where the destination switches are located. This enables the engineer setting up the switching pattern to work undisturbed, and there is a minimum of traffic to cause dust. It is obviously of the utmost importance that no setting-up errors or operational faults occur a[t] point.

(5.6) Clock System

Accurate time is of prime importance in all broadcastin[g] particularly so in the External Services, where only secon[d] allowed for sender, aerial and wavelength changes. A ba[...] operated master-clock system is used.

Slave dials throughout the building are operated from o[...] other of the same two master clocks as the a.s.u. The s[...] is divided into loops comprising a number of clocks in [...] Each loop is operated by a relay. The relays are connec[...] series with the contacts on the master clock. This limi[...] number of clocks which will fail if a fault develops in a cl[...] the wiring and also limits the voltage across the master-[...] contacts. There is automatic correction of the master cloc[...] Greenwich time signal.

In general, technical areas are provided with second[...] other areas with half-minute clocks.

(6) STUDIO EQUIPMENT

The studio equipments at Bush House consist of a new [...] design, a B.B.C. wartime design (suitably modified) and a[...] mercial unit (also suitably modified).

(6.1) Type B Marks I and II Studio Equipments

The new B.B.C. studio equipments used at Bush House [...] two types—Mark I in talks studios and Mark II in ge[...] purpose studios. Each comprises a control desk, an appa[...] bay and a power-supply cabinet. The Mark II also inclu[...] source and cue light selection cabinet.

The control desk is an assembly of sub-panels of the [...] size, with one exception, and desks can therefore be ma[...] meet various requirements by selecting suitable sub-panels[...]

The apparatus bay is equipped with standard amplifier[...] associated mains units, according to the requirements of the[...]

The source and cue-selection cabinet enables studio [...] phones and outside sources to be plugged to specific chann[...] the desk and cue lights in the studio to be controlled by s[...] cue keys on the desk.

The power-supply cabinet controls technical power, li[...] and 50-volt supplies for the suite.

The facilities provided by control desks vary from one [...] to another, but a typical arrangement using the Mark I [...] ment is:

(*a*) Three microphone channels.
(*b*) One gramophone channel.
(*c*) Three outside sources with cue programme and talkba[...]
(*d*) One incoming contribution with clean feed.
(*e*) Main fader.
(*f*) Visual and aural monitoring.
(*g*) Remote start for recording/reproducing tape machines.
(*h*) Talkback to studio.
(*i*) Cue keys.

With the Mark II equipment a typical arrangement is:

(*a*) Eight microphone channels, individually switchable to [...] of two groups.
(*b*) One independent microphone channel.
(*c*) Two gramophone channels, one of which can be switc[...] enable effects records to be played into the studio loudspeak[...]
(*d*) Two outside sources with cue programme, talkback an[...] feed to one of them.
(*e*) Adjustable echo on any channel.
(*f*) Two group faders.
(*g*) Echo fader.
(*h*) Main fader.
(*i*) Public address system, switchable to selected channels.
(*j*) Special effects units which can be switched in or [...] circuit on two channels.
(*k*) As (*f*) to (*i*) for Mark I.

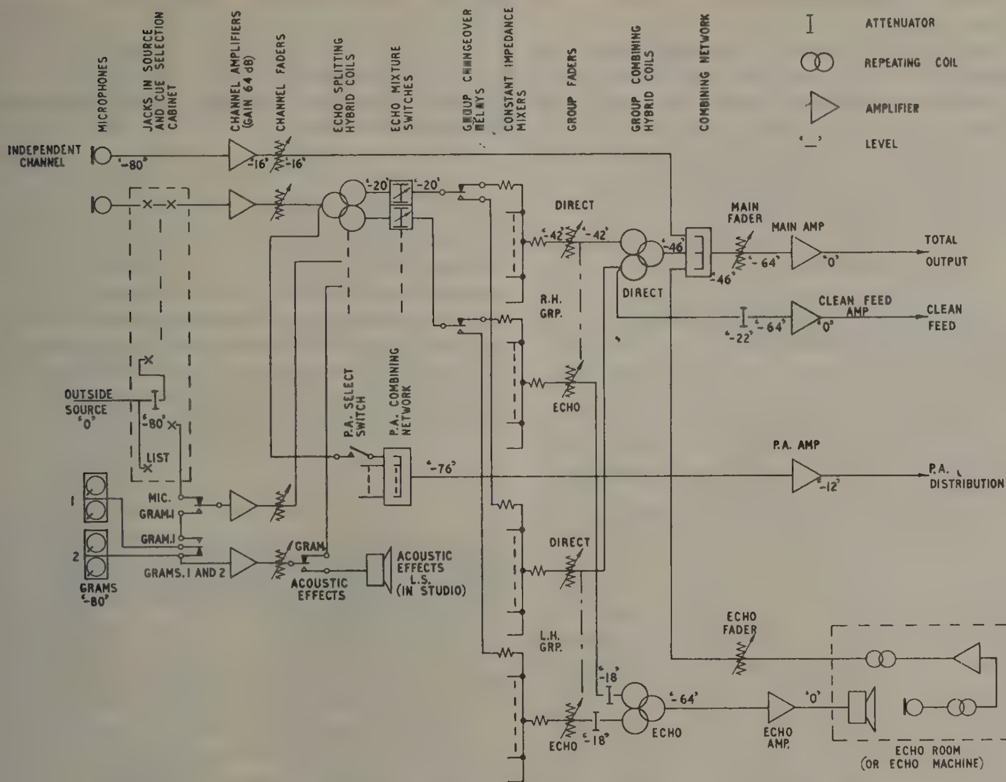

Fig. 4.—Type B Mark II simplified programme block schematic.

The main variations between individual desks are the numbers microphone channels and outside sources. Fig. 4 is a simplid programme block schematic of the Mark II studio equipment. Many of the above facilities are self-explanatory, but a few es will clarify others.

Clean feed' is the total output of a desk, less specified channels. the case of such programmes as 'Two-way Quizzes', the dback to the remote point, where it may be reproduced on dspeakers, must not include that point's own contribution. ernatively, a programme without local announcements may required for the addition of special announcements in foreign guages, or for recording for future use.

Talkback is available in studios on a loudspeaker during earsals, but is restricted to headphones during transmission, ess all microphones are faded out.

The special effects units enable deliberate distortion to be oduced on the channels into which they are switched.

n a studio suite with a Mark I desk, each microphone is tied a specific channel. In the Mark II desk, all microphones selected to the required channels and there may be more rophone points in the studio than are used for a particular gramme.

n both Marks, mixing is carried out after the microphone plifier and at constant impedance. The level is thus indedent of the number of sources faded up.

(6.2) Modification of Existing War-Time Equipment

When the first studios were planned for Bush House the ne considerations were availability of equipment and speed installation. A simple design was produced, based on the of outside broadcast equipment, using low-level mixing. From time to time, various facilities were added. When it was decided to retain some of this equipment, further modifications were necessary to make the best use of the facilities provided by the new control room. An outside source was added, with cue programme and talkback facilities. No provision was made for clean feed.

(6.3) Commercial Equipment

Before the type B equipment was designed, some talks studios were equipped with a commercially available control desk. This does not provide all the facilities of a type B Mark I, but it is adequate for some studios. It employs constant-impedance high-level mixing and has switchable outside sources. Modifications were made to provide cue programme and talkback.

(7) RECORDING AND REPRODUCING EQUIPMENT

(7.1) Disc

The equipment used in two of the three recording channels is some of the American equipment supplied during the war and extensively modified to use B.B.C.-type cutter heads with negative feedback and suction swarf-removal equipment added. A pick-up is fitted to each machine for direct monitoring. An associated bay of equipment contains the recording amplifiers, limiter, monitor amplifier and radius compensation units.

The equipment in the third channel is a B.B.C. type D recorder, consisting of two machines with a recording amplifier in the pedestal of each and a central linking console in which are pick-up pre-amplifiers, line, loudspeaker and peak programme meter amplifiers. On the top of the console are control panels, a peak programme meter and monitoring and change-over keys.

The recording amplifiers, with a maximum output of 75 VA, feed B.B.C.-type cutter heads.

In the reproducing channels, two B.B.C.-designed twin-speed reproducing desks (78 or $33\frac{1}{3}$ r.p.m.) with a bay containing a linking/fader panel and monitoring facilities are used. The turntables use direct rim-drives and facilities are available for remote control of the motors. The pre-amplifier and a prefade listen amplifier for each machine are housed in the desk itself, but the common line amplifier and the main prefade listen amplifier are mounted on the bay.

(7.2) Tape

Four recording/reproducing channels have been provided, each with two tape recorders. Remotely-controlled start and stop for reproduction is available and a delay relay is inserted in the reproducing line to obviate the possibility of the run-up of the machine being radiated.

A linking console contains a control panel, relay panel, peak programme meter amplifier and telephones. The control panel is provided with visual and aural monitoring, recording and reproducing change-over keys and recording fader.

In addition to recording and reproducing, considerable use is made of these channels for editing purposes, means being provided on the consoles for dubbing from one machine to another.

A comparatively new departure in B.B.C. practice is the installation of two central tape rooms. Each has ten channels to the control room, a channel in this context comprising send and receive music lines, a telephone line and various signalling circuits. Channels and machines appear as a 15-pin socket on a distribution panel in the machine room and may be double-ended to any desired machine or machines in parallel. Relays, attenuators, monitoring keys and amplifiers are accommodated in pairs in small consoles between pairs of machines, and on each console there is a recording/reproducing linking panel with change-over keys, recording fader, peak programme meter and remote-start facilities. The inputs and outputs of the linking panels also appear as 15-pin sockets on the distribution panel and may therefore be plugged in circuit as desired. Remote control from the destination for reproduction and from the source for recording is available in both rooms.

Comprehensive monitoring facilities are provided in an adjacent cubicle, separated by a soundproof observation window. All machines and the outgoing music line of any channel may be monitored and the telephone lines of all channels are available. In addition, there is a lamp indicator showing the state of each channel and machine, whether set for recording or reproducing, for remote or manual start, and whether running or not.

Fifteen-wire tie lines are provided between the two central rooms for emergencies and also between central tape room No. 1 and the special-projects room.

Three editing/dubbing suites are each equipped with three tape machines, a linking console and a disc recording channel. Various types of disc reproducing desks are also installed to enable extracts from existing recordings to be combined with current productions. Recording, playback and telephone lines to the control room are available, so that the suites may be used as normal disc or tape recording channels.

Although not strictly an editing suite, the quality check room may be included in this group. It is equipped with a disc recording channel, a tape recorder and various types of disc reproducers. Its function is to check recordings of which the quality may be in doubt and, mainly by means of frequency-response correcting devices, to make them suitable for use.

The special-projects recording room, while again not strictly an editing room, is also included here for convenience. In this suite, there are three tape recorders and a disc recording cha together with a linking and distribution bay. There are sending channels to each of two adjacent studio cubicles. in a central tape room, any tape machine or the disc ch may be 'double-ended' to any of the sending channels, wi without a linking panel.

(8) POWER SUPPLY EQUIPMENT

Power for all areas is normally drawn from the public su mains. The arrangements in the south-east wing and c block are similar. In each case, two independent supplie provided from the mains substation and the load is di between them with interconnection facilities.

In the event of a power failure, essential services, includi technical equipment, reduced lighting in technical areas emergency lighting on stairs, etc., are maintained by three start Diesel-driven generating sets with ratings of 10 kVA (si phase), 22 kVA (3-phase) and 67 kVA (3-phase) respectively

In order that the timing of programmes shall not be by excessive variations in mains frequency, a frequency-stab supply is available in all recording and reproducing room which the drive motors of the recorders or reproducers m switched. This supply is derived from a 12·5 kVA m alternator which is started manually whenever the mains quency differs from 50 c/s by more than ± 1 c/s or whenev standby Diesel generator is in use. The frequency stabil this supply is better than $\pm 0{\cdot}05$ c/s from no-load to full lo

Auxiliary power for the operation of the selector switche lamps in the various indicators and the relay panels in all nical areas is derived from a 50-volt 250 Ah battery, wh float-charged by one of two chargers, either of which is ca of delivering the full load required by all the connected appa This load could amount to a short-period peak of about 60 The normal current is from 7 to 15 amp.

(9) INSTALLATION

(9.1) Timetable

An outline specification of requirements was issued b External Services to the Building, Designs, and Plannin Installation Departments in June, 1954. The first step w produce an estimate of the total cost. In March, 1955, decided to proceed with the project and detailed plannin started.

It was desirable to relinquish 200, Oxford Street as so possible, to avoid revenue expenditure on two premises. was, however, a great deal of design work to be done manufacturing information could be issued to the Equi Department. There was also a considerable amount of bu work to be carried out. A timetable was drawn up, car dovetailing building, design and manufacturing periods. agreed that the Oxford Street premises should be vacated end of 1957, and a target date of September, 1957, was fix completion of installation and testing in the control-room and new studios. Installation commenced in May, 1956 testing in June, 1957. The first network was transfer September, 1957. By November, all networks had been ferred and the new control room suite and studios were operational. The original studios were still working th the old control room.

The transfer of the original studios from the old to th control room proceeded slowly throughout 1958, sinc studios could only be released from service in pairs. T control room finally ceased to operate in October, 1958.

(9.2) Types of Cables and Methods of Running

Until the Bush House installation, the general practice

C. had been to use lead-covered cables between areas and nter-bay wiring, the latter being accommodated in floor s.

ne pros and cons of overhead trays and floor ducts were idered for Bush House. Overhead trays enable cables to un at different heights and facilitate changes, if necessary, later date. They are also more flexible, since it is easier to ll additional trays than to form new floor ducts in a com-d area. On the other hand, entry to a desk is at floor level a bunch of cables dropping down from above would be ghtly. Furthermore, the laying of cables had to commence re all bays of equipment were available, and this precluded g the trays to the tops of the bays. To hang the trays from ceiling would have necessitated strengthening the floor e—an undesirable and expensive feature. It was decided the advantages of more space and greater flexibility out-hed the disadvantages and an overhead tray system was gned, the trays, in one or more tiers, being supported on ghts erected between adjacent bays (Fig. 5). The parts

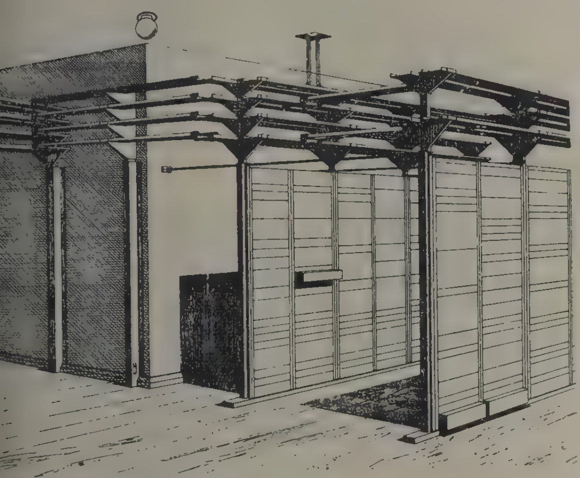

Fig. 5.—Overhead tray system.

l together by sliding over one another. The area behind desks in the control room formed a 'surface' floor duct, and cables feeding the desks were taken from the overhead trays oor level in one large trunking fixed against a wall.

aving decided on an overhead system, weight precluded the of lead-covered cables. In any case, the cost of plastic-red cables, including laying, is less than that of lead-covered this type was used throughout. Microphone circuits were in 6½ lb screened pairs. The long tie lines from the control n to the distant studios and recording areas were run in multi-cored cables with the pairs individually screened. t other circuits were run in 6½ lb multi-cored cables, with an all screen.

ne magnitude of the installation can be realized from the roximate quantities of cable used (excluding the manufacture ays, desks and other apparatus), which are:

1 pair	8 miles
5 pairs	5 miles
10 pairs	23 miles
25 pairs	7 miles

he cabling to the main distribution frame, which has nearly terminal blocks averaging 200 tags each, received very careful ance planning.

(9.3) Testing

Units which were complete in themselves were tested in the Equipment Department test room. Many units, such as relay panels, switching panels, etc., were only parts of a complex arrangement of equipment which could not be set up for complete testing before installation at site. The same applied to assembled bays and desks.

Much of the design was new and could only be tested, at the design stage, on a small scale. It will thus be appreciated that site testing was a major item. That three months sufficed is a tribute to the careful attention to detail and planning in the design, manufacturing and installation stages.

(9.4) Transfer of Services from Oxford Street and Bush House Old Control Room

Until August, 1957, all networks were fed into the a.s.u. in the old control room, either by Post Office lines from Oxford Street or from the old continuity suites at Bush House.

As a network was transferred from Oxford Street to the centre block, its continuity output (from the new control room) was fed into the old control room over a tie line and thence into the a.s.u. The Post Office lines to the transmitters were still connected to the a.s.u. in the old control room. By November, 1957, the a.s.u., still in the old control room, was fed by all networks from either the original continuity suites in Bush House or from those in the centre block via the tie lines.

In the meantime, the Post Office had provided T-joints to the outgoing cables which were connected to the new control room through U-link panels with the links removed.

Two operations now proceeded in parallel:

(*a*) The original studios were taken out of service, modified, connected to the new continuity suites and then handled in the same way as those transferred from Oxford Street.

(*b*) The chain switching in the old control room was transferred from the a.s.u. to the original X and Y bays and operated manually, whilst the a.s.u. was transferred to the new control room.

The a.s.u. inputs were wired to the outputs of all continuity suites without disconnecting them from the X and Y bays. All chains were thus available simultaneously on the X and Y bays and the a.s.u., the latter being isolated from the T'd lines by the omission of the U-links in the new control room. When all was ready and fully tested, the change-over between the manual and auto chain switching only required the removal of U-links in the old control room and the insertion of the links in the new control room. Finally, the continuity outputs to the X and Y bays were disconnected and the Post Office cut away the cables to the U-link panels in the old control room.

Apart from dealing with the remaining original studios and carrying out a few minor modifications, found desirable in service, this completed the work of centralizing the External Services in Bush House.

(10) PRACTICAL EXPERIENCE WITH THE INSTALLATION IN SERVICE

The introduction of centralized working in the new control room allowed a direct saving of 12% to be made in engineering staff. Indirectly, the way was also prepared so that this saving was increased to 14% when studio and recording activities were integrated. Experience in the operation of the new control room has allowed proposals for a further reduction, bringing the total saving up to about 17%, despite increasing commitments in the meantime.

The reliability of the equipment as a whole has been excellent and similar installations have been confidently planned for the Broadcasting House Extension and the Television Centre. In

its early days, the a.s.u. suffered from a certain amount of contact trouble, traced to the ingress of dirt from the atmosphere. The present location of the equipment appears to be satisfactory in this respect. Some selection of valves for the first stages of microphone amplifiers has proved to be necessary to reduce noise.

The new equipment has greatly improved the general standards of ease and convenience of operation, as well as providing additional facilities which can be summarized as follows:

Studios.—Additional facilities are provided on the type-B desks.

Control Room.—Compared with previous standards, the new control room is operationally much more convenient because of the juxtaposition of the miscellaneous switching position and the engineering manual exchange, together with completely cordless operation of these units.

The operational staff have the assistance of a continuous visual indication of the actual loading and can carry out, from one position, stage-by-stage visual and aural monitoring of the programme chains for continuity and quality checking purposes.

Control and monitoring positions have been provided.

Recall facilities are provided on extended telephone circuits and conference facilities are also available on the e.m.x., avoiding unnecessary repetition of service messages, etc.

The majority of the amplifiers can be replaced rapidly by the switchable spares provided on the racks. Replacement c amplifiers themselves is a simple and speedy matter, since are held in the panels by single quick-operated fasteners connected by plugs and sockets.

Continuity Positions.—With the exception of the three tinuity suites, these are now concentrated in the control area instead of being scattered about the various buildings. new equipment provides much greater speed in the selecti alternative sources.

There is continuous visual indication of the source select each channel and the state of readiness of remote recordin reproducing channels.

Additional features include the provision of Big Ben wi preselection and remote starting facilities for recording as as reproducing channels. On control desks, certain compo such as the main fader, can be by-passed and spare amplifie be switched to replace faulty units.

(11) ACKNOWLEDGMENTS

The authors wish to thank Mr. H. Wilkinson, formerly B.B.C., and their colleagues in the Building, External Se Broadcasting, and Planning and Installation Departmen assistance in the preparation of the paper. They are indebted to the Chief Engineer of the B.B.C. for permissi publish it.

DISCUSSION BEFORE THE ELECTRONICS AND COMMUNICATIONS SECTION, 2ND MAY, 19

Mr. F. C. McLean: One could easily get the impression these days that nothing much happened other than in television. The paper shows that a great deal of thought is required to make other jobs go also.

External broadcasting has to be particularly good to meet the competition from other users of the short-wave spectrum. The only way to retain our audience is to ensure that the broadcast is good, well-timed and of fine quality. That is the only attraction that we can offer. Others have the advantage of better geographical situation, greater power, etc. It is efforts such as are described which make all this possible.

It is interesting to note that the power of the transmitters which this programme of installation feeds is more than twice that of Home, Light and Third put together, as is the output figure of 80 hours. To feed so many transmitters from so many studios calls for very careful timing, and the automatic switch is a very elegant solution to that problem. What can be done to override the automatic switching if some international event takes place and it is necessary at short notice to change the whole schedule?

The substitution of plastic- for lead-covered cables is most interesting. The saving in cost and the greater cleanliness are extremely valuable, and perhaps also more attractive. Have we paid a price for this, for example in increased trouble from crosstalk? What is the audio level at which programmes are distributed and switched? This can also have an effect on the crosstalk situation.

There is no reference to the B.B.C. programme and news-collecting activities. The great attraction of our programmes throughout the world is that our news bulletins are up-to-date and accurate, and it would have been interesting to have some information on the teleprinter network and other facilities for bringing news into Bush House and getting it quickly into the hands of the appropriate people.

Reference has been made to both disc and tape recording. What is the proportion of each, and is there a movement towards the greater use of tape? What is the tape speed used in the studio and in mobile operations?

Mr. F. Williams: I speak as a user from Home Sound b casting. We also are in the throes of planning a new c room on similar lines and are changing to an above-g daylight area. This is an improvement in working cond which will undoubtedly have a psychologically beneficial so far as the operators are concerned.

The staff saving referred to is primarily achieved, I thi the concentration of the controls at a central operating des the authors have not stressed sufficiently the other fac available at the desk, which make this saving possible, pa larly the ability to monitor at various points in any chain source to destination. The operation of the desk is not switching sources to destinations. It is a matter of bein quickly to locate a fault and to take corrective action. operator is able to do all this without moving from his fewer staff will be required than would otherwise be nece

On the question of switching, the system adopted is t switching sources to channels, and then channels to destina In other words, to get a source-to-destination involve operations. The purpose, I gather, is to avoid the large n of uniselectors that would otherwise be needed. What full effect of this saving? It is surely cumbersome to h undertake two switching operations when one will suffice. many more uniselectors would be involved in a direct-swi system, and would this obviate the need for the au indicating board?

In the Home Sound services there are only three net which operate almost continuously, as opposed to a very larger number in the External services, operating, wi exception of one or two, somewhat spasmodically. The number of networks undoubtedly justifies a central repro room, but in Home Sound we have come to the conclusio it is better to associate these reproducing machines wi services into which the programmes are to be reproduce to locate the reproducing channels in the respective con rooms. Even in the best-run services, errors, either hun caused by faulty equipment, will occur, and speed of comm tion between the operators at the continuity and repro

ions is of paramount importance if prompt and effective n is to be taken.

Home Sound we have found a rather unique solution to roblem of overhead trays, which have been put under the . This was achieved by placing the selectors and the main bution frame on the two floors immediately below the ol room, under the equipment with which they work. This very short runs between these units and unsightly overhead in the control room itself are avoided.

r. **R. D. Petrie:** Two further reasons for the rejection of and-jack (patch cord) switching were the absence of clearly ble displays of programme traffic conditions and the lity to connect a large number of ancillary circuits simul- usly with the connection of programme circuits. This disability could, of course, be avoided by the use of multiple and sockets arranged in the form of a cross-hatch or e of connectors. This is done in some control rooms on Continent. An installation of the size described in the paper, ever, would need selection panels many square feet in area, moreover, the panels would have to be repeated at every hing position. The programme source multiple, too, d have to wander round the various operating areas. some programmes very short items follow one another in succession, and programme sources must be preselected for sequential fading in and out of circuit. This means multiple socket panels must therefore be increased in area n amount equal to their basic size for each additional pre- tion required.

e argument for some form of remote control switching is fore overwhelming. The authors refer to the adequacy of switching for small installations, and to give some scale e statement I estimate that for equivalent facilities a motor lector system becomes more economical than relays at a hing capacity of about 30 sources to six destinations, and the r the installation, the more economical becomes the use of r uniselectors. All switching is carried out at zero level the measured crosstalk over the complete switching chain een correctly terminated adjacent sources was 70dB at .

is probably worth mentioning, in reference to the engineer- nanual exchange, that when a source-to-route programme it has been completed, telephonic communication is also lished between the terminal points without the intervention e e.m.x. operator. I think it worth stressing the importance e 400-point monitoring system associated with the e.m.x., t this position the operator, on receiving a telephoned query ding a circuit fault, can immediately scan the relevant it under his control to locate the site of the fault.

r. **H. Snell:** The authors say that they have considered hing by plug and jack and relay, as well as by uniselectors, ave they considered a method that is now coming into use— onic switching? The Richmond telephone exchange is operated in this way, using cold-cathode tubes.

r. **E. W. Hayes:** It is interesting to consider what is implied e broadcast transmitting station by the demand for rapid hing of programmes and their routing. First, we have dvantage that the adoption of the automatic switching unit e transmitting station makes it unnecessary for the trans- r operator to be able to identify languages. Switching s are reduced by this automation, because operators need dentify similar languages.

the other hand, the need for rapid switching introduces problems. It becomes necessary at times, for instance, to on a 100 kW transmitter in the permitted 20 sec. The mitter is commonly standing by with only the filaments d and has to go through all the processes from that point on, including any lining up that is possible, in the 20 sec allowed. This is commonly called by transmitter staff a 'crash start'.

In the new transmitters now being installed, provision is made for this 20 sec change of programme or programme routing in substantially the same way as the X and Y arrangement described by the authors, by having two independent output stages to each of the two transmitters. Wave changes are completed in 20 sec and automatic switching of the aerial in the same period is provided for. We are hoping to extend the automatic switching unit so that the impulse for initiating the wave change on the transmitters will, in fact, be a signal from the automatic switch unit. We envisage as an eventual target that the whole operation may be automatic, from Bush House to the trans- mitting station aerial.

In Section 6.1 reference is made to a device to enable deliberate distortion to be introduced on the channels into which it is switched. This, if used judiciously, might have great value for squeezing out a little more money from an administration for the maintenance of broadcasting plant, but I have no doubt that it has a more mundane application. Will the authors say what this is?

Mr. M. M. Freeland: It is not clear at what point during the 20 sec silence the switching is carried out. Is it at zero, or some time before? If it is carried out before, what precautions are taken to ensure that the announcers have, in fact, stopped talking?

The size given for studios is about 3 000 ft^3, say 20 ft $\times$ 15 ft $\times$ 10 ft. That is, after all, the size of a fairly big living room, yet the type B Mark II equipment has provision for eight microphone channels. I should like to know how they can be used. Even the smallest talks studio has provision for three microphones, which seems over-lavish.

I suggest that the ratio of tape to disc recording will probably be quite high—four to one—yet, although we have disc repro- ducing equipment in the studio, there are no facilities for local tape reproducing.

Has the studio reverberation time of 0·35 sec been chosen as a result of experience in this country on the Home service, or has consideration been given to the distortion—atmospheric etc.—and the adjacent-channel noise that is encountered in overseas broadcasting?

Mr. E. A. Beaumont: The reliability of the equipment obviously depends very much upon the correct functioning of the large number of contacts in the uniselectors, and in Section 10 reference is made to trouble with the a.s.u. in the early days due to the ingress of dirt. This trouble has recurred on several occasions, and there is little doubt that it is due to dirt, as stated. It may be argued that the right way to deal with this situation is to stop the dirt getting in; unfortunately this is not possible in the case of the marking code selectors. The screenwork sur- rounding this part of the equipment must be opened once daily, on average, to allow rearrangement of the combs: there is not the same necessity to open the protective screens round the other switches.

It seems that the design of the actual contacts is unsatisfactory in the presence of dirt, and it takes a skilled maintenance engineer about one hour to clean the contacts of one uniselector switch. Would it not be possible to devise an improved contact? The one that springs to my mind is the hard-line type in which a knife-edge contact travels circumferentially, pushing aside any dirt encountered. In my experience it is much superior to contacts of the area type.

Mr. M. J. L. Pulling: We have had a clear description of the arrangements whereby the programme progresses from its source at Bush House to its destination at a transmitting station. It occurs to me that an error might occur which could cause a

programme to be fed to the wrong transmitter aerial array. I should like to know whether there is any check which permits instant detection of this, or whether one has to wait until a message comes back from a distant part of the earth that someone hoping to receive a particular programme has received a different one?

Perhaps one of our authors could say something about the audience to which these programmes go. I imagine tha these days a large proportion of the audience gets them thr a local relaying station, probably on medium waves, and the organization that picks up the programme direct from B.B.C. operates a special receiving station for that purpose there any evidence that a considerable number of people ov in fact still listen direct on short-wave receivers?

THE AUTHORS' REPLY TO THE ABOVE DISCUSSION

Messrs. F. Axon and **O. H. Barron** (*in reply*): Mr. McLean refers to last-minute changes in the transmission schedule. Over-ride keys are provided for manual operation of the a.s.u. and these also cover fault conditions. Comprehensive monitoring facilities are provided.

All news bulletins broadcast by the External Services are prepared in Bush House. The main sources of news are the press agencies and the B.B.C. Monitoring Station at Caversham; an extensive teleprinter network links these and other sources, such as the Houses of Parliament, to the News Room. After the bulletins have been prepared, they are distributed to the programme sections in other parts of the building by pneumatic tube and teleprinter.

There is a continuing trend towards greater use of tape for recording. At the present time tape claims 80% of the total recorded output, and this is likely to increase in the future. As a general rule, a tape speed of 15 in/sec is used for recordings of music and $7\frac{1}{2}$ in/sec for speech.

We agree with Mr. Williams that the staff saving is largely due to centralization of the controls. The functions of the main operating desk are described in Section 5.2. A direct-switching system for Bush House would require an additional 174 switches (15 bays). The price of this saving is double switching and a limited number (50) of simultaneous connections. In the case of the installation planned for Broadcasting House this limitation is not acceptable. A direct-switching system is to be used, since double switching with over 50 channels does not show a worthwhile saving. An auxiliary indicator will not be necessary. We are glad Mr. Williams liked the relative disposition of bays and overhead trays planned for Broadcasting House.

Mr. Snell refers to electronic switching. The designers considered this and decided that the cost would be much higher than motor uniselector switches and that the crosstalk would be inadequate for broadcast purposes.

Mr. Hayes is correct in his assumption concerning deliberate distortion. The special-effects units control audio-frequency bandwidth, mainly in drama productions, to produce effects such as telephone conversations.

Replying to Mr. Freeland, programme switching takes place 18 sec before zero. In order to avoid being interrupted by the switching operation, announcers maintain silence from zero minus 20 sec to zero minus 15 sec.

Type B studio equipments were designed so that two sizes covered all normal requirements. Usually Mark I is used in studios for talks, discussions etc., and Mark II in larger general-purpose studios. Most of the desks in Bush House are Mark I. Whether it is preferable for reproducing equipment to be or in a central room depends on the type of service. As Williams remarked, local tape-reproducing has been found convenient for Home Services, but central rooms are well s to External Services. Recently, it has been found convenie Bush House for some networks to tape-reproduce local continuity suites, and trolley-mounted machines are used this purpose, supplementing those in the central rooms. reproducing equipment is local to the studios in all B services, and is frequently used for selected portions of for which centralized working would be unsuitable. The s reverberation time of 0·35 sec was based on experience i Home Service and was considered satisfactory for the Ove Services.

We think that Mr. Beaumont is over-stressing a.s.u. tr due to dirt. Records show a total of 12 switching faults o a.s.u. during the six months to June 30th, 1960, and seven d the preceding six months. During the same periods, faul all other switching equipment (348 switches with 800 con each) were 19 and 31, respectively. We would not expec a.s.u. protective covers to be opened daily, since the over panel can cover occasional changes in the schedule. The of switch had to be one already in quantity production, cluding the use of specially shaped contacts. In any case are not convinced that a knife-edge contact would be factory and consider it would damage the surfaces of the contacts, which are very difficult to replace.

Mr. Pulling raises the question of incorrect progra routing. With several programmes running simultaneously risk that one of them may be incorrectly routed is always pr To legislate for this, the programme radiated by each transm is checked at frequent intervals against the operational sch by the B.B.C. Measurement and Technical Receiving Stati Tatsfield.

With the growth of medium-wave broadcasting overse number of listeners who receive their programmes by re on the increase. Nevertheless, only a small proportion potential audience is within reach of a medium-wave relay so direct listening on short waves is still predominant. Rece surveys and correspondence with listeners confirm this.

Mr. Petrie's contribution includes answers to some points raised by Messrs. McLean and Williams.

For further information about the Empire Service, see a by HAYES, L. W., and MACLARTY, B. N.: 'The Empire S Broadcasting Station at Daventry', *Journal I.E.E.*, 193 p. 321.

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HENRI DE FRANCE COLOUR TELEVISION SYSTEM

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ie paper was first received 29th June, 1959, *and in revised form* 11*th February*, 1960. *It was published in April*, 1960, *and was read before the* ELECTRONICS AND COMMUNICATIONS SECTION 27*th April*, 1960.)

SUMMARY

e paper describes the work done by the Cie Française de Télévision .T.) on the Henri de France system of colour television. First it es the objectives of the team of research workers who have carried his development. It then stresses the fundamental aspects and the al features of the system and describes the encoding and decoding . Finally, it gives the practical results and theoretical aspects on ase of transmission of the coded signals.

(1) INTRODUCTION

is well known that a colour picture may be defined by three ponents of information and that a 'black and white' picture ls only one component of information. It is also well known the three components of colour information may be treated rently from each other during transmission. It is desirable ransmit the luminance component—which in practice is a icular combination of three primary (red, green and blue) ponents over a wide bandwidth, but the remaining two linear binations of the three primary components may be transed over a narrower bandwidth. Thus the wide-band inforion component is luminance and the two other components espond jointly to the chrominance.

rom the aspect of the general theory of communication the blem is to determine how these three components of informa- should be transmitted and how they should be encoded in r to ensure a specific quality while occupying as small a uency spectrum as possible. To these data it is necessary to an essential point of exploitation, namely that of triple patibility, i.e.

(*a*) Chrominance information must be easily extracted from the mposite signal, leaving only the luminance information for onochrome reception of a colour picture on a 'black and white' ceiver.

(*b*) 'Black and white' pictures must be capable of being received a colour-television receiver.

(*c*) The frequency band occupied by the colour picture informan must be no wider than that occupied by the black-and-white nal.

arting from this basis, the development of the Henri de ce system has been concentrated on the simplification of the ving circuits and on reducing the difficulties associated with long-distance transmission of colour pictures. The simty of receivers ensures, in general, ease of adjustment and ility of performance, and this fully justifies the first objective. realization of a colour-television picture transmission is an nsive matter; it is therefore important to ensure widespread smission and, concomitantly, to take advantage of the nsive network of microwave links established for the transion of black-and-white pictures. In practice, this demands ncoding of colour television signals suited to transmission litions identical with those which secure good quality for k-and-white television. The search for this special encoding the second objective of the development.

nsieur Chaste is with the Cie Télégraphie Sans Fil, and Monsieur Cassagne is he Cie Française de Télévision.

(2) PRINCIPAL CHARACTERISTICS OF THE HENRI DE FRANCE SYSTEM

In known systems of colour television, when the bandwidth of the chrominance signals is reduced the horizontal colour definition is lowered to an extent which is not perceptible to the eye while vertical definition remains practically unchanged. This fact seems rather abnormal when one considers that, in all standards, bandwidth and the number of lines must be related to achieve a reasonable balance between the horizontal and vertical definitions. The first important aspect of the Henri de France system is that it takes advantage of the fact that there is little point in supplying the eye with colour information which has more definition in the vertical direction than in the horizontal direction; with the horizontal colour definition reduced it is also possible, without disadvantage, to supply the added chrominance information in alternate lines in the vertical direction. The application of this idea gives the system its second, and no less important, characteristic. It is clear that the simultaneous transmission of the three signal components of colour information over a given frequency band raises greater risks of interaction between these signals than would be the case if only two signals were transmitted simultaneously over the same frequency band. The sequential transmission of chrominance information—the means employed in the Henri de France system to ensure the correct relationship between the vertical and horizontal colour definitions, makes it possible to transmit only two signal components simultaneously.

The luminance signal is transmitted over a wide band for all scanning lines, in accordance with the usual rules for black-and-white television, but each of the complementary chrominance signal components is transmitted, over a narrow band, in alternate lines. The chrominance information which is not transmitted differs very little from that corresponding to the preceding line, which has already been transmitted. At the receiver, the information is considered as being identical, since the eye cannot appreciate the difference. This therefore provides three simultaneous signals while only two have actually been transmitted. For any one line the three signals are luminance, the chrominance information transmitted during that line and the chrominance information transmitted during the preceding line. This last signal is made available by the inclusion of a delay line* which repeats the signal after a delay of a line period. At any instant the chrominance signal transmitted follows two paths: one takes it directly to the decoding matrix before the receiving cathode-ray tube, while the other makes it available for the subsequent line via the delay line.

At the transmitter the two chrominance signals, multiplexed in time, are frequency multiplexed with the luminance signal so that they may be transmitted at the same time as the latter. The luminance information keeps its normal place in the frequency spectrum. A sub-carrier allows the two chrominance signals to be transposed (see Fig. 1) and is situated at the end of the video spectrum with which it is interlaced; it can be transmitted with

* The delay line used for decoding is a rather special element of colour television receivers in the Henri de France principle; in order to produce simple and inexpensive receivers, special attention has been given to reducing its price and size, and Section 4 gives full information on this point.

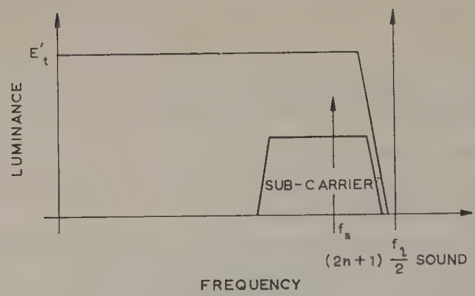

Fig. 1.—Frequency spectrum of composite signal.

symmetrical or asymmetrical sidebands. In order to reduce its effect on the luminance signal, its frequency is chosen as an odd multiple of half the line frequency, as in the N.T.S.C. system; with the same object, other precautions are taken and are described in Section 8.2. The arrangement obtained in this way is compatible, and the black-and-white picture can be extracted from the composite signal by a simple filter. Conversely, since the principle of constant luminance is observed, as in the N.T.S.C. system, a colour receiver reproduces a black-and-white picture faithfully.

Lastly, stress should be laid on the fact that the chrominance sub-carrier is at all times modulated only by a single signal. This secures all the advantages of a single modulation. For example, if the sub-carrier is amplitude-modulated, there is no need to be concerned about spurious phase modulation. If the sub-carrier is phase or frequency modulated, then the behaviour of the amplitude of the sub-carrier has no significance.

In short, it can be said that the Henri de France system is characterized by

(*a*) Sequential transmission of chrominance information line by line.
(*b*) Simultaneous transmission of two components of colour information instead of three.
(*c*) Single signal modulation of the chrominance sub-carrier.
(*d*) Storage of the chrominance information at the receiver from one line to the next.

(3) GENERAL PRINCIPLE OF AN ENCODER-DECODER UNIT

(3.1) Composition of the Transmitted Signal

Let E'_t be the input voltage of a transmitter modulator in the analysis of a colour picture. We can write:

$$E'_t = E'_y + \begin{bmatrix} E_{c1} \\ \text{or} \\ E_{c2} \end{bmatrix} \cos 2\pi f_s t + S'_y$$

where
$$E'_y = \alpha E'_G + \beta E'_R + \gamma E'_B$$
$$E_{c1} = E_{c0}[a + b(\alpha' E_G + \beta' E_R + \gamma' E_B)]$$
$$E_{c2} = E_{c0}[a + c(\alpha'' E_G + \beta'' E_R + \gamma'' E_B)]$$

E'_y is the luminance voltage with γ-correction, and E_{c1} and E_{c2} are the chrominance sequential voltages. S'_y is the synchronizing signal carrying the line and frame synchronizing signals, S_{yL}, of a conventional black-and-white picture, and pulses $S_{yc1}\cos(2\pi f_s t)$ or $S_{yc2}\cos(2\pi f_s t)$ of sub-carrier frequency f_s, depending on the line. These pulses allow E_{c1} and E_{c2} to be identified: their width is approximately the same as that of the synchronizing pulses of the signal S_{yL} and they are situated on the back porch. E'_R, E'_B, and E'_G are the γ-corrected voltages corresponding to the respective red, blue and green signals which result from the analysis. E_R, E_B and E_G are the low-frequency components of these voltages, which suffice for defining the chrominance si α, β, γ, a, b, c, and E_{c0} are coefficients.

The factor E_{c1} or E_{c2} to be taken in consideration i equation which gives E'_t is a function of the analysed line c sponding to the picture point under consideration, and ch from one line to the next as a result of time-division multipl previously described.

Section 8.1 gives a set of numerical values correspondin particular case of number of lines, for sub-carrier value linear combinations for luminance and chrominance, etc., a as the description of the composite signal obtained in the mission of a colour test pattern.

(3.2) Encoder

Fig. 2 is the block diagram of an encoder. This unit pro the composite signal E'_t from the camera voltages E'_G, E E'_B and also the synchronizing voltages and the voltages sub-carrier frequency which it produces itself. The three si E'_G, E'_R and E'_B, issuing from some analysing device are ap to a matrix $\alpha\beta\gamma$ supplying the luminance signal by linear bination of the three signals. After the synchronizing s have been added, the signal E'_y is applied to a delay line w function is to bring together in phase the luminance chrominance signals. The signal is then passed to a chromi mixer which also superimposes the modulated sub-carrier.

The two other matrices, $\alpha'\beta'\gamma'$ and $\alpha''\beta''\gamma''$, produce band chrominance signals E'_{c1} and E'_{c2} which are applied electronic switch controlled at line frequency by the syn nizing-pulse generator. A filter follows the output o electronic switch and reduces the bandwidth of the chromi signals. These narrow-band sequential signals then driv sub-carrier modulator. This sub-carrier is not sent continu to the modulator to be modulated by the chrominance si It is premodulated in order to be only partially present c back porch of the black-and-white picture signal and for which differ from one line to the next. This allows th chrominance signals to be separated.

It can be seen from this description that the networks are simple and without critical adjustment. In particula relatively low value of the switching frequency will be note

(3.3) Decoder

Fig. 3 shows the block diagram of that unit whose funct to extract from the composite signal the wide-band lumi information and the three components of narrow-band ch nance information, red, green and blue. More exactl decoder's function is to reconstitute physically the three ch nance signals of hybrid composition. The 'low frequency' of these signals are separated and correspond to the signa E_G and E_B. The 'high frequency' part, or 'mixed hig common to all the E_R, E_G and E_B signals obtained fro luminance signals. By superimposing these three pr impressions, the human eye then recovers the luminanc chrominance of a picture.

The composite signal E'_t passed to the decoder input i subdivided, by filtering, into two signals—the luminance E the chrominance E_{c1} or E_{c2} transmitted by the sub-carrier. us examine in detail the use of the two chrominance signa and E_{c2}. The first operation to carry out is their identific since either is liable to appear at any instant; this is achiev the pulses S_{yc1} and S_{yc2}, which control the switching gen at the right instant. After identification, two sequential s must be converted to two simultaneous signals. In the e tion of the general principles it was seen that the eye is inse to amplitude variations of E_{c1} or E_{c2} from one line to the In order to effect the sequential-to-simultaneous conve

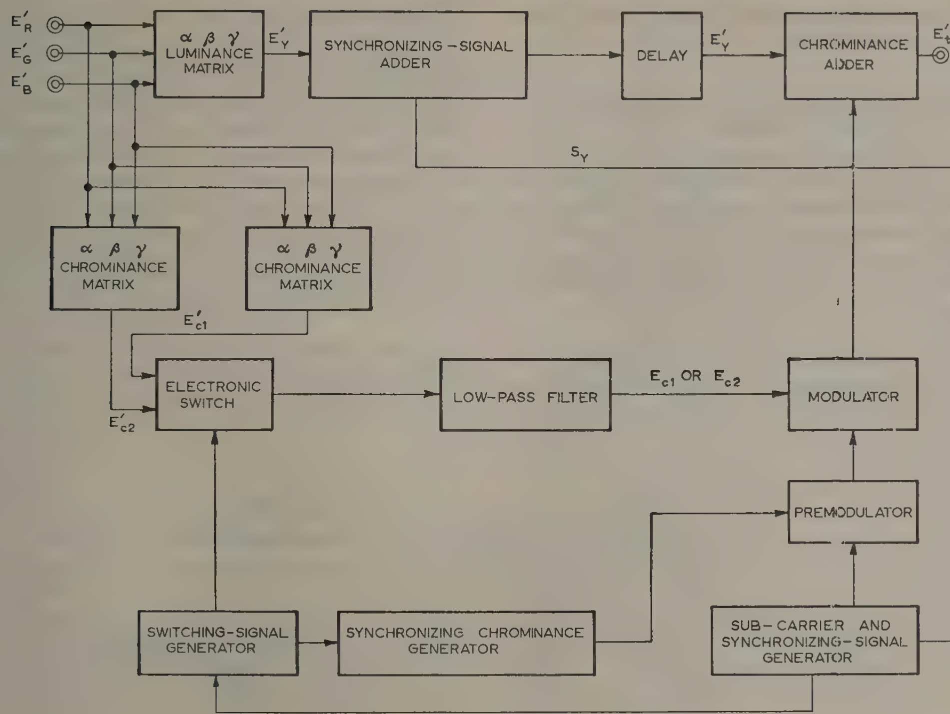

Fig. 2.—Block diagram of encoder.

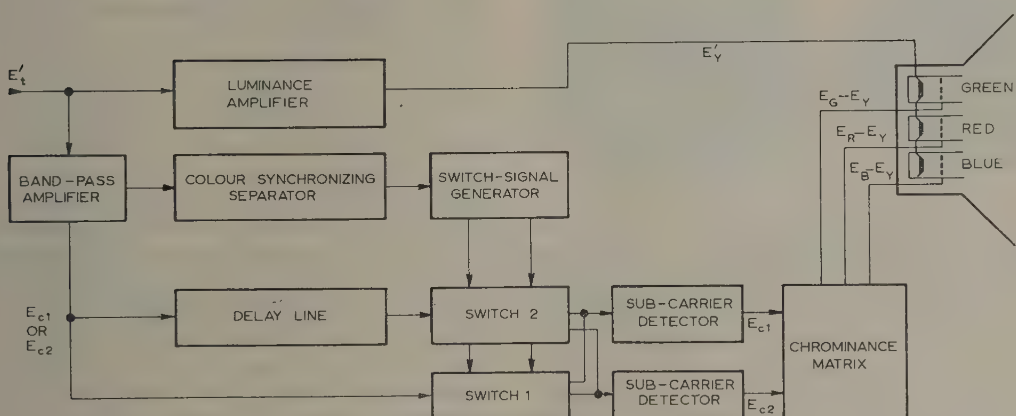

Fig. 3.—Block diagram of decoder.

ystematic use is made of the chrominance information, E_{c1} or $_{c2}$, available at a given moment and that, E_{c2} or E_{c1}, trans-itted during the preceding line. The sequential signals are ansmitted over two channels, one direct and the other with a ansmission delay equal to the duration of one line. At the utput of these two channels two signals are simultaneously vailable, namely E_{c1} and E_{c2}. One output provides in succes-on E_{c1}, E_{c2}, E_{c1}, E_{c2}, . . . and the other provides E_{c2}, E_{c1}, $_{c2}$, E_{c1} . . . All that is needed is to switch these two outputs in ynchronism, at line speed, to two new channels in order to btain two simultaneous and continuous voltages E_{c1} and E_{c2}. hese operations are effected by switches 1 and 2.

The twin outputs of switches 1 and 2 feed two detectors which rectify the sub-carrier envelope, i.e. reproduce the two signals E_{c1} and E_{c2} and not transposed signals.

A chrominance matrix then effects the transformation of (E_{c1}, E_{c2}) into $(E_R - E_Y, E_G - E_Y, E_B - E_Y)$. It will be noted that E_Y contains only the low-frequency components of the chrominance signal E'_Y. Finally, the three voltages $E_R - E_Y$, $E_G - E_Y$, and $E_B - E_Y$ are applied to each grid of a 3-colour tube with three guns of which the three cathodes are fed in parallel by E'_y. Thus each gun is controlled by a signal related to a single primary colour. For red, for instance, this signal is the sum of $E_R - E_Y$ and of E'_Y, i.e. low-frequency components

properly belonging to the red signal E_R and 'mixed highs' $(E_Y' - E_Y)$ valid for the three colours.

(4) COMMERCIAL RECEIVERS FOR THE HENRI DE FRANCE SYSTEM

The design of a commercial receiver for the reception of signals coded in accordance with the Henri de France system will now be examined, and the following description is largely that for receivers which were used for the C.F.T. laboratory tests. It is clear that the future of colour television is conditioned by the possibility of building simple and inexpensive receivers, and this has been the aim of the C.F.T., as shown in Fig. 4.

This diagram shows all the usual parts of a black-and-whi or a colour television receiver. A high-frequency and a intermediate-frequency amplifier bring the received signal to level suitable for detection. It is only beyond the detector that th colour receiver begins to have special features.

The valve V_0, which belongs to the normal chain of an television receiver, not only amplifies the luminance signal bu also isolates the sub-carrier modulated by the chrominanc signals. Its anode tuned circuit filters out this sub-carrie which is then passed to the input of the circuits necessary fo the reception of colour pictures. Another circuit, in all respec similar to that of a black-and-white receiver, supplies to th cathode-ray the luminance, scanning and synchronizing signa

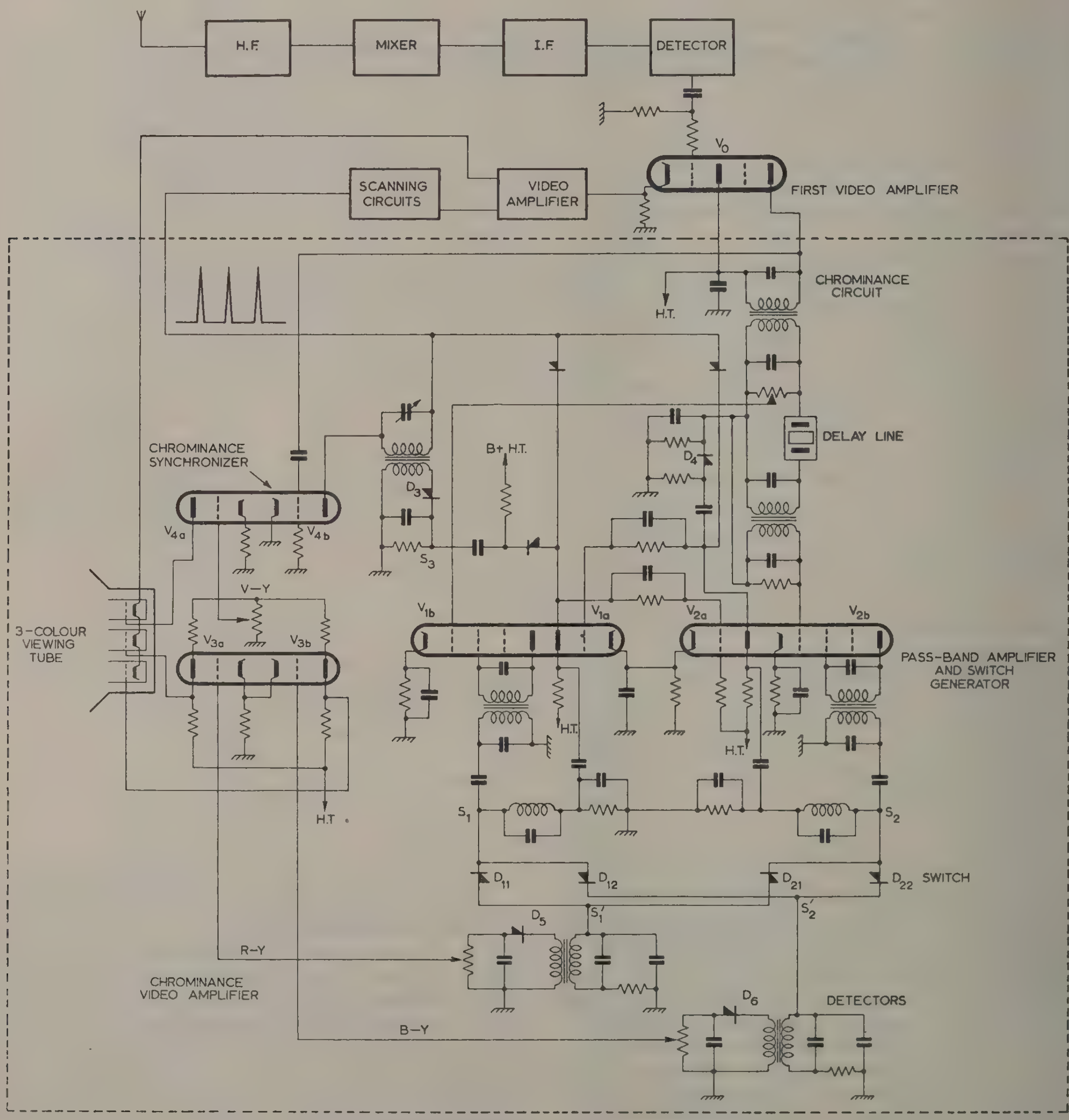

Fig. 4.—Commercial receiver with chrominance circuits.

t also supplies to the 'chrominance' section of the receiver ne-synchronizing signals whose purpose will be described later.

The sub-carrier, filtered by the anode circuit of V_0, drives in arallel the grids of V_{1b} and V_{2b}, the drive to the grid of V_{1b} eing obtained directly from the secondary of the anode circuit f V_0. The drive to the grid of V_{2b} is obtained through a delay ne providing a delay of one line (64 microsec in the 625-line tandard). At any instant, V_{1b} and V_{2b} thus amplify the hrominance signals of two different lines.

By the principle of the system, each of these signals, corre-ponding to two successive lines, belongs to one of the chromi-ance components of information to be transmitted. In the eceiver built these components are respectively $R - Y$ (red, less uminance) and $B - Y$ (blue, less luminance).

The electronic switch mentioned in the previous Section here ses diodes. By blocking or unblocking diodes D_{11}, D_{12}, D_{21} nd D_{22}, it switches the outputs S_1 and S_2 over to the outputs S_1' and S_2'. Thus by judicious synchronization, the output S_1' an always reconstitute a sub-carrier modulated by $R - Y$ and he output S_2' a sub-carrier modulated by $B - Y$. For example, at a given instant the line is being analysed for $R - Y$, the utput S_1' is fed directly with $R - Y$ by S_1 through diode D_{11}. The output S_2' is fed with $B - Y$ by S_2 through D_{22}. The utput $B - Y$ is also fed by the signal corresponding to the ne previously analysed for $B - Y$.

V_3 then amplifies the two signals $R - Y$ and $B - Y$, and by coupling between its two anodes produces the combination $G - Y$ amplified by V_{4a}. The signals $R - Y$, $B - Y$ and $G - Y$ re then available and are applied to the grids of the 3-colour ube, the three cathodes of which are also fed with the luminance ignal.

Returning now to the line-sequential switching, the blocking nd unblocking of the diodes D_{11}, D_{12}, D_{21} and D_{22} is con-rolled by signals from the bistable multivibrator inserted between he two anodes of V_{1a} and V_{2a}. This multivibrator is controlled n the one hand by the line-synchronizing signals taken from he conventional part of the receiver, and on the other by pulses vhich are always produced every other line at the secondary of he anode circuit of V_{4b}.

The line-synchronizing pulses cause a change-over at every ine. The pulses from V_{4b} ensure that this change-over occurs n the proper phase, for the multivibrator can be triggered only y the simultaneous presence of the line-synchronization pulses nd of the pulses from V_{4b}.

The pulses which appear in every other line are produced as ollows. V_{4b} is unblocked only during the beginning of the ack porch by a pulse originating in the sweep circuits. From Fig. 10 it is easily seen that, during this unblocking period, V_{4b} s able to pass the sub-carrier only if it is present, i.e. once every wo lines. Then there is, after detection by D_3, only one pulse every second line at S_3. The chrominance signal transmitted, $R - Y$ or $B - Y$, can thus be recognized without ambiguity.

V_{4b} is also used as a 'colour killer'. In the absence of a hrominance signal there is no pulse at S_3. The multivibrator driven only by the line-synchronization pulses does not turn over and the detection arrangement D_4 produces a zero voltage while biasing positively the grids of V_{1b} and V_{2b} when the multivibrator is in action. Thus, in the absence of a chromi-nance signal, the two chromatic video amplifiers are blocked.

(4.1) Prototype Delay Line

This part of the receiver has been the subject of important development work aimed at reducing its dimensions and lowering its cost. This work was deemed necessary by C.F.T. right from the start and has led to some extremely interesting results.

During the earlier trials it was first necessary to verify the principles of the Henri de France system before the engineering problems could be studied, so that the first delay line used was a simple HH2500 cable. This coaxial line consists of a flexible mandrel of polythene loaded with a magnetic powder, around which the central conductor is wound in order to produce the necessary line inductance. An external metal shield provides a line capacitance suitable for producing a delay of 2 microsec/m; thus 32 m of cable were needed to produce the delay of 64 microsec required for a 625-line system. Although the cable diameter was small, such a delay line was immediately considered too large; moreover it had various technical and economic drawbacks.

(4.2) Second Type of Delay Line

The second line used was of the coaxial type with the external conductor forming a solenoid whose characteristics had been determined to make the phase/frequency curve linear. The experimental model had a characteristic impedance of 3 kilohms and a delay of 22 microsec/m; thus 2·9 m were required to give the requisite 64 microsec delay. This line was still too large and its cost too high. Although it is not considered that a final result has been attained, the latest lines built show that it is already possible to build receivers in which the delay line is of only secondary importance in cost and size.

(4.3) Latest Designs of Delay Line

The latest designs are multiple-reflection quartz lines. Trans-ducers are ferro-electric cells of barium titanate which use the propagation of longitudinal sinusoidal waves in a solid medium

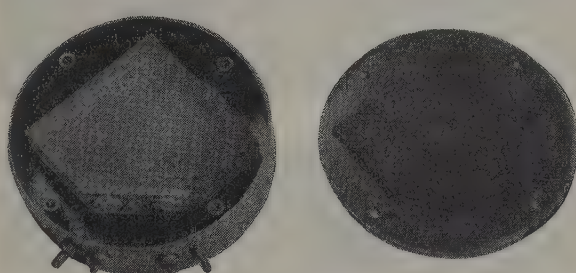

Fig. 5.—Delay line.

of fused silica. Fig. 5 shows the external appearance and the internal construction of these lines. Their main characteristics are as follows:

Delay.—The 64 microsec delay required is obtained structurally with a precision of 0·1 microsec; but since the temperature coefficient of this line is 10^{-4} per deg C, it will be seen that the delay may vary by 0·192 microsec for a temperature range of 30°C. The possible overall difference between the theoretical delay and that obtained in practice is therefore of the order of 0·2 microsec, and this has been verified by experiment (see Section 5.1).

Pass Band.—The amplitude/frequency curve for this line is shown in Fig. 6, the 3 dB points being 1·34 Mc/s apart. The tests mentioned in Section 5.2 were carried out using the pass band symmetrically, i.e. limiting the bandwidth of the chromaticity video signal to about 700 kc/s. A check has also been made regarding the possibility of shifting the sub-carrier by placing it at 900 kc/s from the lower 3 dB point and so at 440 kc/s from the upper 3 dB point. It is, however, possible to widen the pass band of such delay lines and to conform to future standards in respect of the bandwidth of chrominance signals.

Spurious echoes are 26 dB below the useful signal.

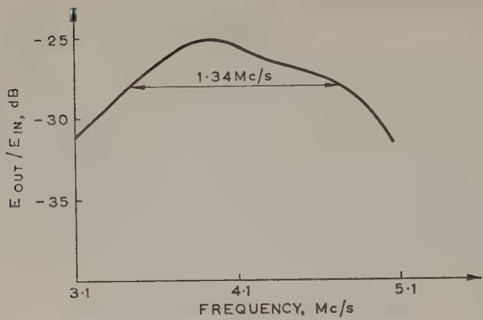

Fig. 6.—Characteristic of delay line.

Attenuation.—The delay line introduces an average attenuation of 26·5 dB.

Transient response.—Fig. 7 shows the input and output signals of the delay line for transmission of a colour-bar test pattern. The good overall fidelity of transmission through the delay line should be noted.

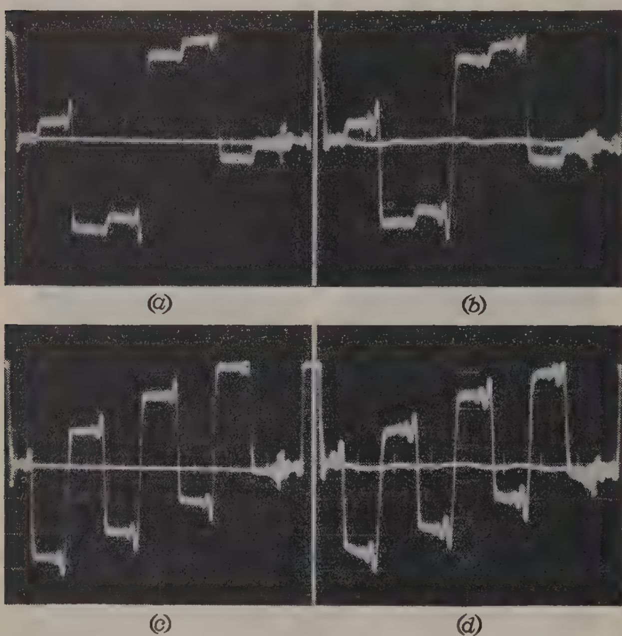

Fig. 7.—Transient responses.

(*a*) Input signal for $R - Y$.
(*b*) Output signal for $R - Y$.
(*c*) Input signal for $B - Y$.
(*d*) Output signal for $B - Y$.

The complete delay line is 1 in high and 3 in in diameter; it costs approximately £6.

Other work is in hand in order to improve certain characteristics and to reduce the price still further, but it is already quite clear from the latest results that the delay-line problem has already become of secondary importance from both the technical and economic points of view.

(5) TRANSMISSION OF COLOUR TELEVISION PICTURES

The arrangement shown diagrammatically in Fig. 8 has enabled C.F.T. to carry out many experiments and demonstrations. This block diagram shows that the colour pictures can

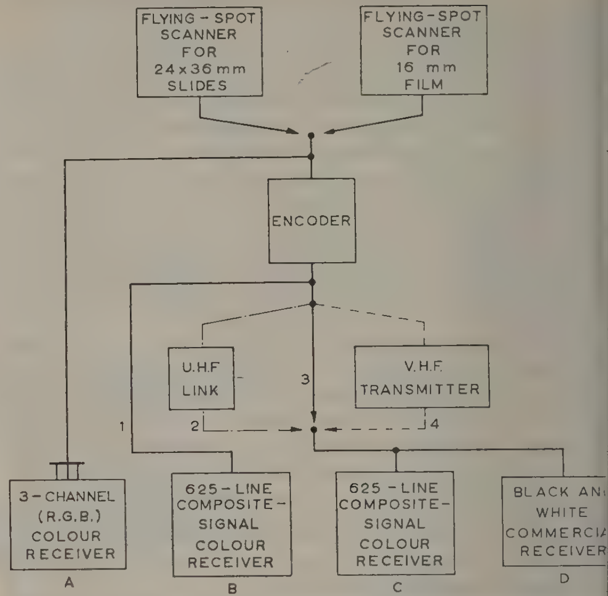

Fig. 8.—Block diagram of apparatus for experimental transmissior

be analysed by two flying-spot scanners, one for 24 mm × 26 mr slides, and the other for 16 mm film. The three signals from th analyser are transmitted, on the one hand to a wide-band receiv with three separate channels, and on the other to an encoder.

The encoder output signal has four possible channel Channels 1 and 3 allow of direct decoding, with reproductio of the colour pictures on an RCA 3-colour tube. Channel provides means for introducing a microwave link in the tran mission, and channel 4 allows transmission via a v.h.f. tran mitter with its associated receiver, free-space propagation bein simulated by means of a variable attenuator. The radio link used more particularly for the study of the effect of accident variations of phase or amplitude, as well as the study of pertu bations caused by noise; the v.h.f. transmitter is used for th study of interference. It is also seen that a black-and-whi receiver enables compatibility to be assessed and that channel is used only for comparison with channels 2 and 4.

It is not proposed to extend the paper to include a detaile description of all the experiments carried out, but the mo striking results will be given for the appreciation of the featur of the system. In the Introduction, stress was laid on the impo tance which C.F.T. attached to the ease of transmission c colour pictures. It is essentially from this viewpoint that th experimental results* and the theoretical points given below a of value.

We shall examine in turn the sensitivity of colour pictures t the three main classes of possible disturbances, namely accident phase variations, accidental amplitude variations, and noise an interference.

(5.1) Effect of Phase Variation

It is under the influence of sub-carrier phase variations th the system shows its most remarkable quality, and this is becaus the chromatic information is not related to the phase. Onl

* The tests were carried out in the following way: a number of pictures were di played in turn on the two 3-colour receivers A and C, fitted with RCA tubes. Tl two receivers were adjusted so as to exhibit two substantially equivalent pictures whe transmitting a half-shade monochrome test pattern and a colour-bar pattern. Tl maximum admissible variations for such or such parameter were determined by tl appearance on the coded picture of colour bands or spurious structures. Clearl the degradation occasioned would often have been imperceptible without the eleme of comparison.

the possibility of a phase shift of colour signals with respect to the original phase determines an upper limit to the acceptable phase variations. It was shown experimentally that a shift of the chrominance information signal of 0·2 microsec with respect to the luminance information signal fixed the acceptable limit. This delay corresponds to a shift of 300° at the frequency of the sub-carriers used (4·1 Mc/s).

The phase of the sub-carrier can vary with luminance, and this is one of the defects which it is most difficult to correct in micro-wave links. It is therefore especially important for a colour television system to be practically independent of the differential phase of the equipment. (Even a long series of radio repeaters could hardly show a differential phase of 300°.)

As in the other systems, non-linearity of the phase/frequency curve in the neighbourhood of the sub-carrier causes an incorrect reproduction of the chromatic transients. But these defects are less marked because they are related to a modulation of a single type, which, moreover, is an amplitude modulation. In particular, they cause no intermodulation between the chrominance signals, even if these are modulated with asymmetrical sidebands.

(5.2) Effects of Amplitude Variations

Since the chrominance information is carried exclusively by an amplitude modulation, about the same susceptibilities are found, for amplitude variations, as in the N.T.S.C. case. It is useful to point out that this amplitude modulation, as in the N.T.S.C. case, corresponds to a single chrominance information component.

The tests carried out have shown that, under the conditions set out above, the tolerance is 1 dB for the differential gain. A higher differential gain causes a hue shift towards green or purple, depending on its sign.

A 1 dB variation in the luminance/chrominance ratio is also admissible. Since the system functions on the principle of constant luminance, variation of the relative ratios merely produces a variation of colour saturation. Much larger variations are ineffective on the synchronization of the chrominance signals, for the few cycles of sub-carrier which occur on the back porch carry only an ‘all or nothing’ signal which is relatively insensitive to level variations.

(5.3) Effects of Noise Interference

The introduction of uniform noise in the transmission spectrum produces no colour shift, only colour desaturation. The degradation of the colour picture is more substantial than that of the monochrome picture with the same signal/noise ratio. Comparisons are very difficult to make, but it would seem that the threshold of noise visibility on the colour picture is about 10 dB below that on the black-and-white picture. An important improvement can be expected from tests which will be referred to at the end of this Section.

The effect of colour-picture interference patterns, of course, increases as the interference frequency approaches the sub-carrier frequency, as shown by Table 1.

In particular it will be noted that, when the frequency of an interfering signal is exactly that of the sub-carrier, the interference pattern becomes visible for a level 13 dB below that which makes it visible on a black-and-white picture. It should be noted that this value corresponds to a visibility threshold and not to an acceptability threshold, which is much higher and very difficult to establish.

To conclude this Section some reference will be made to tests (which show promise of interesting results) at present in hand in the C.F.T. laboratories. The foregoing results and considerations show that the Henri de France system is notably insensitive to sub-carrier phase variations and that it is fairly comparable

Table 1

EFFECT OF INTERFERENCE

Frequency of interfering signal	Level of the interference on the occurrence of interference patterns (with respect to the carrier)
Mc/s	dB
f_0 (carrier frequency)	−38
$f_0 + 3{\cdot}1$	−41
$f_0 + 3{\cdot}6$	−49
$f_0 + 4{\cdot}1$ (sub-carrier frequency)	−51
$f_0 + 4{\cdot}6$	−47
$f_0 + 5{\cdot}1$	−41

to other systems as regards the effect of amplitude variations and noise.

In order to improve the performance of the system from these two aspects C.F.T. have undertaken tests on frequency modulation of the sub-carrier instead of amplitude modulation. It should be noted that frequency modulation does not make coded signals corresponding to large colour patches more sensitive to phase variations than amplitude modulation: the sub-carrier frequency takes a fixed value which characterizes the hue, and the phase is not a carrier of information. At the most, frequency modulation renders the transmission of colour transients more nearly faithful. Against this, as has been confirmed experimentally, it allows of very large tolerances in the differential gain, on the amplitude/frequency law, and in the constancy of the luminance/chrominance ratio. It also brings about a very substantial improvement in protection from noise and interference.

(6) CONCLUSION

Undertaken with very modest facilities, the development of the Henri de France system has very quickly produced results which show notable advantages in comparison with the N.T.S.C. system.

If laboratories of international status could take up the study of improvements to be made to this system, it would certainly be possible to attain the objectives of permitting the construction of inexpensive receivers, easily adjusted and with stable performance, and simplifying programme exchange.

Once these objectives were attained, C.F.T. would feel that it had usefully served the cause of colour television.

(7) ACKNOWLEDGMENTS

The authors wish to express their thanks to M. Henri de France, creative leader of the C.F.T. team. They also wish to thank M. Polonsky, Technical Manager of the C.S.F. Television Group for the extensive advice he gave in the drafting of the paper.

(8) APPENDICES

(8.1) Composition of the Transmitted Signal in the Special Case of a 625-Line Picture

The principal aspects of the 625-line system used for the tests were as follows (levels given in relative values):

Luminance	$E'_y = 1$ (peak-to-peak)
Synchronizing signal for luminance information	$S_{YL} = 0{\cdot}43$ [peak-to-peak (i.e. $\dfrac{S_{YL}}{E'_Y + S_{YL}} = 0{\cdot}31$)]

Synchronizing signal for chrominance information	$S_{YC} = 0{\cdot}25$ (peak-to-peak)
Linear combinations transmitted as chrominance signals	$b(\alpha' E_G + \beta' E_R + \gamma' E_B) = 1{\cdot}43(E_R - E_Y) = 1{\cdot}43\,(0{\cdot}7E_R - 0{\cdot}59E_G - 0{\cdot}11E_B)$ $c(\alpha'' E_G + \beta'' E_R + \gamma'' E_B) = 1{\cdot}12(E_B - E_Y) = 1{\cdot}12\,(-0{\cdot}3E_R - 0{\cdot}59E_G + 0{\cdot}89E_R)$
Chrominance signal	$E_c = 0{\cdot}125 \left\{1 + \begin{bmatrix} 1{\cdot}43(E_R - E_Y) \\ \text{or} \\ 1{\cdot}12(E_B - E_Y) \end{bmatrix}\right\}$
Sub-carrier frequency . .	4·1 Mc/s
Delay introduced in one chrominance channel on decoding	64 microsec
Duration of sub-carrier elimination for suppressing chrominance	6·5 microsec for one line, 8 microsec for the next
Luminance signal	$E'_Y = (0{\cdot}59E'_G + 0{\cdot}3E'_R + 0{\cdot}11E'_B)$

The particular values of the coefficients α, β and γ correspond to the use of the reproduction primaries whose co-ordinates are, on the C.I.É. system of axes,

Red	$x = 0{\cdot}67,\ y = 0{\cdot}33$
Blue	$x = 0{\cdot}14,\ y = 0{\cdot}08$
Green	$x = 0{\cdot}21,\ y = 0{\cdot}71$

The variables E_G, E_R, E_B, E'_G, E'_R, E'_B are between 0 and 1, and the variables $E_R - E_Y$ and $E_B - E_Y$ are between -1 and $+1$.

It is thus possible to state the value of the composite signal obtained during the transmission of the colour-bar test pattern regularly used for adjusting colour transmission systems. It will be noted that this pattern contains, in addition to black and white, which are uncoloured values, the three fundamental completely saturated shades red, green and blue, and their three complementary colours: blue-green (cyan), violet (magenta) and yellow.

Table 2 gives the set of numerical values which the colour variables take during such a transmission.

The waveform of the composite signal is shown in Fig. 9. It will be seen that, in this particular case, the composite signal has a peak-to-peak amplitude 8·75% greater than that of the black-and-white signal, and that the maximum incursion o the sub-carrier beyond black has an amplitude equal to 10% o the black-and-white signal, i.e. 33% of the synchronizing signal

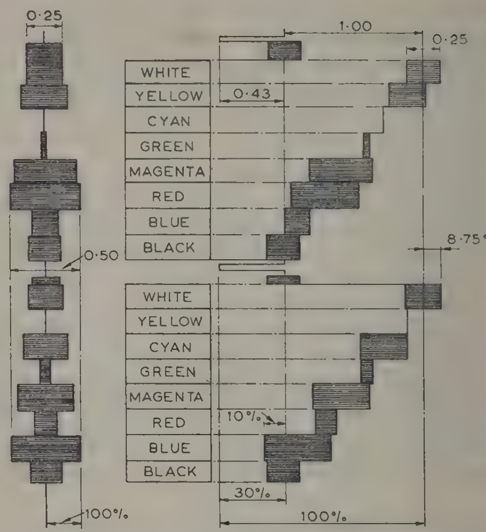

Fig. 9.—Sub-carrier signal patterns produced on cathode-ray oscillograph.

(8.2) Precautions taken to Reduce the Visibility of the Sub-Carrier Interference Patterns

The interlacing of the luminance and chrominance spectr naturally causes intermodulation of these signals. This applie to all systems of colour television when the total spectrum widtl is equal to that of the luminance spectrum alone. We shall se out the precautions which have been taken to reduce the effect of that intermodulation.

First, considering the presence of the sub-carrier and of it sidebands in the luminance signal, the choice of a frequenc which is an odd multiple of half the line frequency, known a 'offset', makes it possible to obtain low-visibility spots such a are shown in Fig. 10(*a*). There occurs a partial plus or minu compensation effect between the brilliance values due to th positive or negative half-waves of the sub-carrier, taking int account the practical equality of the signal in two successiv lines or in two corresponding lines in two successive frames

Table 2

RELATIVE VALUES OF COLOUR VARIABLES DURING TRANSMISSION

	White	Yellow	Cyan	Green	Magenta	Red	Blue	Black
E'_R	1	1	0	0	1	1	0	0
E'_B	1	0	1	0	1	0	1	0
E'_G	1	1	1	1	0	0	0	0
E'_y	1·00	0·88	0·70	0·59	0·41	0·30	0·11	0
$E_R - E_y$	0	+0·12	−0·70	−0·59	+0·59	+0·70	−0·11	0
$E_B - E_y$	0	−0·88	+0·30	−0·59	+0·59	−0·30	+0·89	0
E_{c1}	0·125	0·145	0	0·02	0·23	0·25	0·10	0·125
E_{c2}	0·125	0	0·170	0·04	0·21	0·08	0·25	0·125
E'_t (peak)	1·125	1·03	0·87	0·63	0·64	0·55	0·37	0·125

$$E' = E'_y + E_c \cos 2\pi f_s t$$
$$E'_y = 0{\cdot}59E'_G + 0{\cdot}30E'_R + 0{\cdot}11E_B$$
$$E_c = 0{\cdot}125\left[1 + \begin{matrix} 1{\cdot}43(E_r - E_y) \\ \text{or} \\ 1{\cdot}12(E_B - E_y) \end{matrix}\right] = 0{\cdot}125 + \begin{bmatrix} 0{\cdot}18(E_r - E_y) \\ \text{or} \\ 0{\cdot}14(E_B - E_y) \end{bmatrix}$$

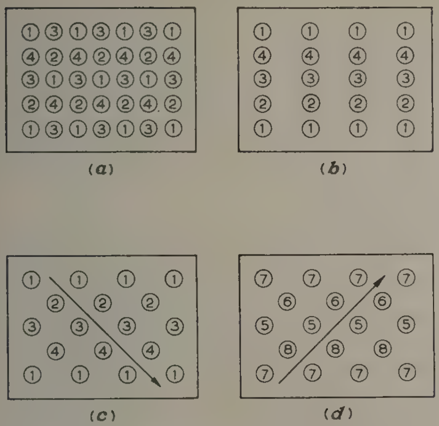

Fig. 10.—Patterns of an 'offset' sub-carrier.

Every point shows the positive peak of the sub-carrier and each numeral gives the number of the field.
(a) N.T.S.C. case.
(b) Continuous recurrence of sequential multiplex.
(c), (d) Discontinuous recurrence of sequential multiplex.

However, in the case of the Henri de France system, two successive lines actually bring almost similar luminance information; but this is no longer true for the modulated sub-carrier signal, which carries two information components, $R - Y$ and $B - Y$, which are usually different.

Fig. 10(*b*) shows the spotting which appears when no precaution is taken in the time-multiplexing recurrence frequency, in the extreme case for which either $R - Y$ or $B - Y$ is assumed to be zero and the other a maximum. For example, it will be seen from Fig. 10(*b*) that the spots, 3, of the first line have disappeared while those of the third line have remained; this is because the chromatic information of the first line of the third frame is zero while that of its second line (i.e. of the third line of the Figure) is assumed to be a maximum. The Figure shows the substantial increase in visibility made possible by the appearance of columns and a less dense structure of the spotting. Moreover, it shows that the order of succession of the frames, 1, 2, 3 and 4, leads to an upward drift [as is also the case in Fig. 10(*a*)].

These defects can be reduced by altering the multiplexing recurrence at every frame and the phase of the sub-carrier. The method used is as follows:

(*a*) At the end of each of frames 1–4 a discontinuity is introduced in the succession of the chromatic information components $R - Y$ and $B - Y$. This, for instance, leads to exploring the first line of frame $n + 1$ as the last line of frame n for $R - Y$. This first line of the $n + 1$ frame is thus explored for $R - Y$, while it would have been explored for $B - Y$ with a continuous multiplexing recurrence. Spotting then has the appearance shown in Fig. 10(*c*). It has a diagonal arrangement, which reduces its visibility. The order of succession of frames 1–4 causes downward drift whose bad effect is attenuated by a second type of discontinuity.

(*b*) In contrast to what happens at the end of frames 1–4, no modification is made to the normal succession $R - Y$, $B - Y$, $R - Y$, etc., signals at the end of frames 5–8. But the phase of the sub-carrier is reversed at the end of these frames, which produces the spotting shown in Fig. 10(*d*). The diagonal arrangement is the same as that in Fig. 10(*c*), but drift now occurs in the upward direction.

In connection with interference patterns, an advantage can be indicated which is inherent in the Henri de France system. As is well known, most commercial receivers show such patterns when, at the transmitter, the picture and mains frequencies are not strictly related. This is always the case in colour television, since the picture frequency is derived from the stable frequency of the sub-carrier crystal in order to reduce the visibility of the sub-carrier. For second-order interlacing, for instance, the picture frequency defined by the sub-carrier crystal may at times differ from the mains half-frequency, which is variable. The interference patterns then become more troublesome as the frequency difference increases.

In the Henri de France system there is no very-narrow-band circuit such as a synchronous detection circuit to demand that the same chromaticity sub-carrier frequency shall always be rigorously maintained. It is therefore perfectly feasible to equip a transmitter with three crystals giving frequencies which are a constant multiple of the mains frequency, e.g.

$$4{\cdot}100\,\text{Mc/s} = 82\,000 \times 50\,\text{c/s}$$
$$4{\cdot}182\,\text{Mc/s} = 82\,000 \times 51\,\text{c/s}$$
$$4{\cdot}018\,\text{Mc/s} = 82\,000 \times 49\,\text{c/s}$$

and automatically to switch the crystals at the transmitter as the mains frequency varies between 49 and 51 c/s.

Under these conditions the sub-carrier difference at the receiver (2%) remains acceptable and the interference patterns are substantially reduced, the difference between the picture frequency and half the mains frequency then being lowered.

DISCUSSION BEFORE THE ELECTRONICS AND COMMUNICATIONS SECTION, 27TH APRIL, 1960

Mr. W. J. Bray: This first demonstration of the transmission of colour television signals over the Eurovision network from Paris to London has been an historic occasion. In view of the long-distance transmission of the signals, it might be of interest to give details of the links used (Fig. A). The link from Paris to the P.T.T. terminal at Tour de Meudon was by temporary microwave equipment. From there, the signals were fed over the P.T.T. permanent microwave link to Loos and thence to Fiennes on the French coast and the Post Office radio station at Tolsford Hill, near Folkestone. From Tolsford Hill the signals were transmitted to Crystal Palace, London, by temporary microwave links set up by the British Broadcasting Corporation. The normal routing for the Eurovision signals is by coaxial cable between Folkestone and London; however, the bandwidth on the existing coaxial cables on this route is 5 Mc/s, whereas the 625-line colour signals require a bandwidth of about 5 Mc/s and it was therefore necessary to set up the temporary microwave links.

From Crystal Palace the signals proceeded by coaxial cables, with a bandwidth of 5 Mc/s, to Broadcasting House, the Post Office television switching centre at Museum Exchange and thence to Gerrard Exchange, near the Institution building. I might mention that The Institution is now connected by permanent coaxial cable link to the national television network, so that in future members will have the advantage of that facility.

The overall performance of the Paris–London link will be of interest. The overall bandwidth was effectively 5 Mc/s, and the signal/noise ratio was 50 dB weighted and 43 dB unweighted. The waveform performance of the overall system was such that the *K*-factor—the rating commonly accepted for assessing the waveform transmission performance of television links—was 5% before waveform correction. Even with this *K*-factor and without additional waveform correction, the quality of the received colour signals was acceptable. However, some additional waveform correction was inserted and the *K*-factor reduced from 5% to about 1%. The linearity of the system was

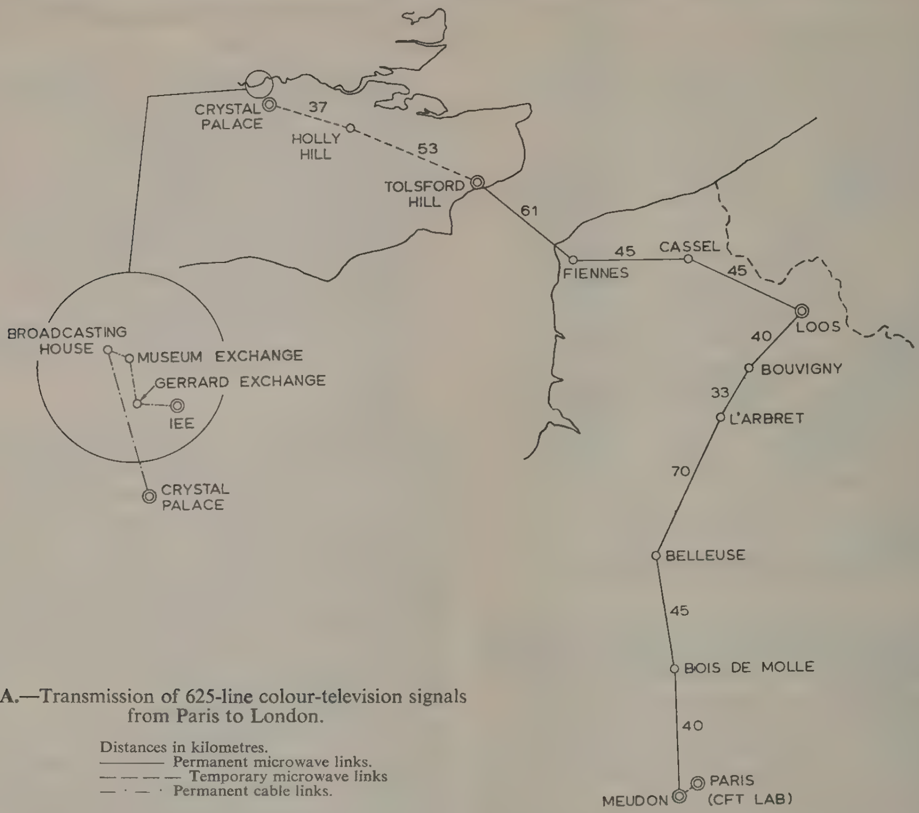

Fig. A.—Transmission of 625-line colour-television signals from Paris to London.

Distances in kilometres.
——— Permanent microwave links.
— — — — Temporary microwave links
— · — · Permanent cable links.

such that the black signals in the standard staircase test pattern were compressed from 100 to 82% and the white signals to 96%, so that the overall system was slightly deficient in respect of linearity.

The signal transmitted from Paris used pre-emphasis of the video signal in which the low-video-frequency components were reduced in amplitude by some 10 or 12 dB before transmission. At the receiving terminal—in this case the microwave-link terminal at Crystal Palace—a de-emphasis network was provided to restore the video signal to its correct proportions. The use of pre-emphasis greatly reduced the impairment of the video colour signal due to differential amplitude and phase distortion on the microwave links by restricting the frequency deviation corresponding to areas of uniform brightness.

Mr. V. J. Cooper: This evening we have witnessed two historic events—our first contact with Europe through colour television, and a demonstration in this country of the Secam system, which the most conservative among us will acknowledge as coming nearer to being a serious competitor of the N.T.S.C. system than anything we have seen so far.

Beginning with the N.T.S.C. system, the authors have recognized two difficulties, namely long-distance transmission and the necessity of using a synchronous detector in the receiver. It is possible that they have exaggerated the importance of these two difficulties, because the Americans are regularly sending N.T.S.C. system colour-television signals over thousands of miles satisfactorily, and currently available equipment designed to transmit 600 speech channels will transmit N.T.S.C. colour signals without significant distortion. We have found the synchronous detector as reliable and robust as any other part of the receiver, and in fringe areas it is common experience to see the colour synchronization holding in conditions of interferenc when the conventional synchronizing is beginning to fad These two alleged difficulties arise because of the double modu lation of the sub-carrier in the N.T.S.C. system, and the firs logical step was, of course, to separate the two modulatio components and permit envelope detection at the receiver and choice of amplitude or frequency modulation. Alternate-lin switching of chrominance has been chosen to effect the separatio

There is also the interesting by-product of the Secam syster of having reduced colour resolution vertically as well as hori zontally. This feature raises two interesting points so far a colour resolution is concerned. The chrominance informatio is delayed for one line in the sequence, i.e. two lines in th make-up of the picture. It is, in fact, out of register with it luminance counterpart. Not only have we low resolution, bu misregistered low resolution. Moreover, a receiver working i a fringe area where horizontal synchronization is not impeccabl will either have very-short-time-constant line-synchronizing per turbations or, if it has a flywheel, it will have long-time-constan perturbations. However, the delay line in the receiver does no know anything about it and will delay the previous line infor mation by a constant amount, disregarding the synchronizin errors. In this case there will also be a horizontal displacemen of colour information relative to the luminance information.

In Section 8.2 the authors refer to the possibility of using 49, 50 and 51 c/s frame frequency. If the conventional line/fiel relation is used it will correspond to a 2% change of lin duration, and this will mean a 2% displacement of chrominanc information for a constant delay line. This is surely unacceptabl

Monsieur R. Sueur (*France; translated by Dr. R. D. A Maurice*): The French Post Office have shown great interest i

this demonstration. The official international organizations have the duty of making recommendations respecting the use of such systems as these, but we should help them by trying to solve the preliminary problems and facilitate implementation of the various recommendations by standardizing transmission characteristics as much as possible. In this way the meetings of the international groups would be shorter: there would be less to do and more leisure to enjoy. We should learn, not only from past successes, but from past failures. France, for her part, has obtained her experience of television on what is now an internationally known standard system, and hopes that there will not be a recurrence in future of the incompatibility of standards experienced in the past.

I am not a television specialist; my main interest is in transmission. It is the engineer who is usually blamed when there is a technical breakdown in national and international programme exchanges. For this reason I ask you, as television engineers, to make our task easier. Black-and-white transmission over great distances is already difficult enough. What will become of colour transmission if differential gain and differential phase, for example, are added to the existing difficulties which we must overcome?

This evening's experiment should be long remembered, for it is the first time that colour television has come to you from the Continent. Moreover, it is an excellent example of international co-operation.

To-day we come to you and say, 'Here is a system in which the signals are very simple and easy to transmit. A few imperfections may still have to be overcome; if that is so, let us work together to complete what has been begun and help develop this unique system which will bring us closer together'. Elsewhere I have seen systems which differ from this: in my opinion, they are not better than it—to the contrary.

Mr. N. N. Parker Smith: It is interesting to note that this system is not arranged so that the sub-carrier vanishes for colours of zero saturation, such as blacks, greys and whites. Several advantages accrue from this property of the N.T.S.C. arrangement. For instance, in the event of a failure in the chrominance signal, the picture seen on the colour set is in black and white. Under this arrangement it would seem to appear in a colour ranging in hue from yellow to cyan which would certainly be less acceptable as an emergency condition of operation. Furthermore, a useful and rapid check may be made by connecting the red, green and blue outputs of the camera in parallel and observing that the sub-carrier is everywhere zero. In the Henri de France system the sub-carrier would appear to have a constant value, which is not so easy to measure.

It would seem that the loss in vertical definition is, in fact, greater than one would expect from the description given in terms of lines n and $n + 1$, because, owing to the 2 : 1 interlace, it is actually the $n + 2$ line which is repeated and the loss in vertical definition would therefore appear to be at least 3 : 1 and not 2 : 1.

Mr. G. B. Townsend: It is undoubtedly very important that we should be able to bring colour-television pictures over great distances, but we must not forget the vital last 20–50 miles from the transmitter to the domestic receiver. The B.B.C. spends a total of £15 000 000 a year on its national television programme. I believe that 6% of this is spent on lines and linkages, which is at least two orders below the national yearly expenditure on new television receivers.

In the N.T.S.C. 405-line receivers which my company manufactures the number of valves in the decoding section can be compared with the similar portion of the Henri de France receiver. I found that the latter has six valves and two diodes less; this must be offset against the cost of the delay line. Is the quoted price of £6 that at which it will be sold to manufacturers following large-scale production, or is it the retail price?

I was surprised to find no mention of gain controls on the various receiver channels. I examined the effect of a 1 dB change in one of the sub-carrier amplitudes and estimated that, in N.T.S.C. terms, it produces a chrominance change on a desaturated bright yellow, approximating to flesh tone, of nearly 30°. Is this the sort of figure one would expect? In comparing our N.T.S.C. receiver with the Secam receiver I am considering a receiver which has automatic chrominance control and splits the chrominance from the luminance immediately after the vision detector. To handle both signals in the same video stage may give rise to trouble, and I wonder whether the authors considered using G − Y instead of B − Y? Would this not have some advantages in regard to the movement of colours when differential gain occurred, flesh tones moving, not in the green–magenta direction, but more in the yellow–blue direction?

I did not understand the suggestion that noise in the a.m. system causes only a desaturation of colours. I would expect a change of hue if there is a perturbation which changes the luminance signal and the one chrominance signal which is being transmitted at that instant.

Mr. S. N. Watson: Like the author, I think that the correct basis on which a colour system should be judged is that of its ability to produce good-quality colour pictures, together with acceptable compatibility. At the present stage of development, a comparison of systems based on cost of receivers is not useful unless the one system shows a major advantage over the other; and since this is certainly not true in comparing the N.T.S.C. and Secam systems, we must look at the fundamental system performance. I shall select two aspects of this, out of the many which might be discussed.

The ability of a colour system faithfully to portray hue is of paramount importance; the authors are of the opinion that a weakness of the N.T.S.C. is the fact that the hue is conveyed by the phase of the colour sub-carrier. While this may be true in the special case of certain f.m. radio-relay links, which in any case can be designed to be perfectly satisfactory, it certainly does not apply to the important part of the transmission system which begins at the transmitting aerial and ends on the viewer's receiver. It is in this sphere—where the equipment is out of skilled hands and where the overwhelming money investment lies—that the N.T.S.C. system is more robust and satisfactory.

My second point is concerned with compatibility: because the average amount of chrominance information is low, the N.T.S.C. colour sub-carrier is low on the average (about 12% of maximum sub-carrier) and zero whenever there is no colour information. On the other hand, the Secam colour sub-carrier will have an average level of 25% of the maximum level of the N.T.S.C. sub-carrier and is accordingly 6 dB stronger than it. Tests in this country have shown that the dot pattern of the N.T.S.C. system is satisfactory, but bearing in mind the gamma of the display tube, doubling the sub-carrier amplitude might well prove most disturbing, particularly since it is strong on such relatively uncoloured and important detail as the human face and figure. The use of frequency modulation for the sub-carrier, precluding as it does any fixed relationship between it and the scanning frequencies, might well make the compatibility of the Secam system unacceptable.

Dr. R. D. A. Maurice: Although one can quite logically say that, if one does not need horizontal resolution beyond a certain value, it need not be given in the vertical direction, it is nevertheless true that the eye does not cut off very sharply, and every little helps. Therefore, although I agree with the authors that it is logical to lose vertical resolution because the N.T.S.C.

system loses it horizontally, I believe that that must be counted against the system as such.

In Section 4.3, although the authors are, in fact, discussing the variability of the delay with temperature, the field frequency must not be allowed to vary either, because the line frequency must remain constant so that the line period stays equal to the delay of the delay line. Therefore, we must run the system asynchronously from the mains as is done with the N.T.S.C. system. A simple calculation based on 50 c/s shows that a $\frac{1}{8}$ c/s change in supply frequency would alter the line scan duration by 0·2 microsec—which is the authors' figure for limit of acceptibility.

Section 5.3 refers to the question of noise. An interesting check can be made in a roundabout way. In the first paragraph of this Section the authors say that 'Comparisons are very difficult to make, but it would seem that the threshold of noise visibility on the colour picture is about 10 dB below that on the black-and-white picture'. Donald Fink, in his book on N.T.S.C. standards, says that the results of a number of tests in the United States gave a figure of 1 dB as expressing the increased vulnerability to random noise compared with monochrome. The B.B.C. made some fringe-area tests comparing a colour slide and a black-and-white slide. When the latter was transmitted no colour burst was sent. The receiver colour equipment was therefore switched off, and it became a monochrome transmission. They found that in fringe areas, at about 100–200 μV/m on Channel 1, the N.T.S.C. system was about 3 dB more vulnerable than monochrome. We may, perhaps, take as an average, the figure of 2 dB. If the Secam system using amplitude modulation is 10 dB less efficient, the net difference between the two systems may be expressed as 8 dB in favour of the N.T.S.C. system; earlier, the authors give the figure of 9 dB. A calculation of the noise on the R–Y and B–Y circuits in the Secam system transmitting a saturated red picture and using amplitude modulation shows that R–Y is 14 dB, and B–Y 11 dB, noisier than on the N.T.S.C. system. We must hope for a considerable improvement from the Secam system using frequency-modulated chrominance.

With regard to vestigial-sideband distortion with amplitude modulation, the Secam system uses a sub-carrier which is situated in the main sideband and there is distortion due to the relative increase of carrier strength, just as there is with the N.T.S.C. system. If the authors' system had been working on positive modulation, the rise in luminance due to the presence of the chrominance signal would have been a mere 8%, which is quite negligible, and the change in hue would have been much less.

I agree that a filter in the monochrome receiver is a good thing if you do not like dots. It is a bad thing if you like panchromatism or orthochromatism.

Mr. L. C. Jesty: With the N.T.S.C. system the synchronous sub-carrier disappears on white and the positive and negative values of the colour signals are determined by sub-carrier phase. With the authors' system and the two-sub-carrier system* zero sub-carrier corresponds to arbitrarily chosen saturated colours (cyan and yellow in Fig. 9). Crosstalk and dot patterns are therefore more obtrusive. Would it not be possible, however, to use single-sideband modulation of a sub-carrier at the top of the luminance spectrum—4·7 Mc/s or higher with an 8 Mc/s channel—thus making crosstalk negligible?

In 1954 we collaborated with the B.B.C. Designs Department in sending colour television signals over the television links in this country. Both N.T.S.C. type and a two-sub-carrier system were used. Serious phase errors were introduced by the links at that time, but with the synchronous sub-carrier system was always possible to adjust to an 'average phase' at th receiving end and obtain a tolerable colour picture. Whit and greys were, of course, always correct. With the two-sub carrier system, although unaffected by the phase errors, th amplitude control of the chrominance signals was extreme critical, and it was very difficult to maintain the 'colour o whites' and the overall tint of the picture. In other word simple amplitude modulation of chrominance has no built-i sense of whiteness.

A small but interesting point is the authors' use of a 'burs on alternate lines. Does this introduce any inter-line flicker o stroboscopic effects on receivers with slow line fly-back?

At present and for some time past the real problem of colou television is to provide an acceptable receiver. Unfortunatel the authors' proposed transmission system involves an expensiv delay line in every individual colour receiver. Although th factory price of this component has been reduced from £5 (1957) to £6, it will have to be reduced very much more t become unimportant.

Mr. G. D. Monteath: Since chrominance information i alternate lines is discarded, one might expect a greater tendenc to form moiré patterns when horizontal stripes appear in th scene. In a 625-line monochrome system coarse moiré patter are formed when the scene contains horizontal stripes separate by about 1/300 of the picture height. These stripes alternate i phase at 25 c/s. In the Secam system, coarse moiré patter (coloured) could be formed when stripes in the scene are separate by 1/150 of the picture height. These patterns should be we marked, since the camera should resolve 150 lines very wel The method of avoiding interference between chrominance an luminance described in the Appendix should cause the coloure patterns to oscillate up and down at $6\frac{1}{4}$ c/s. Have the author observed such an effect?

The effect could be reduced by using delay lines in the studi to combine the chrominance information obtained from pairs o adjacent scanning lines. One scanning line of luminance dela would then be desirable to improve the registration of luminanc and chrominance.

Dr. Maurice has calculated that the chrominance channe should be 11–14 dB noisier in the Secam system than in th N.T.S.C. system. Now chrominance noise in the Secam syster will be correlated in pairs of lines, and this effect might b expected to emphasize its visibility by (at a guess) 2 dB. Thi would make the Secam system subjectively noisier (in relation t chrominance) than the N.T.S.C. system by 13–16 dB.

Mr. P. S. Carnt: In Fig. 4, when a monochrome signal i being received V_1 and V_2 are apparently biased off and the tw signals from the detectors D_5 and D_6 will be zero. If white i being transmitted during a colour signal, these two detecto outputs are not zero. Thus the result will be a green mono chrome picture. I cannot help wondering whether this i intentional.

Mr. T. Kilvington: Section 5.3 indicates that the threshold o noise visibility for the colour picture is about 10 dB below tha for a black-and-white picture. This refers to uniform noise One of the points which the authors have made is that th Henri de France colour signal is easier to transmit over lon distances; but frequency modulation is used in all long-distanc radio links, so that the noise is not of a uniform character but i 'triangular', with a rising spectrum. Have the authors an information concerning the susceptibility of their system t noise of a triangular character?

Table 1 indicates that the level of interference on the occurrenc of patterns for an interfering signal at the carrier frequency f_c is −38 dB relative to the wanted carrier. The figure reportec

* Haantjes, J., and Teer, K.: 'Compatible Colour Television', *Wireless Engineer*, 1956, 33, pp. 3 and 39.

by the C.C.I.R. for an interfering signal at the carrier frequency is −45 dB, and this refers, not to the threshold, but to a level of interference which is regarded as tolerable for a small proportion of the time. I am a little puzzled as to the basis for the figure of −38 dB, and wonder whether the authors can explain it.

THE AUTHORS' REPLY TO THE ABOVE DISCUSSION

Messrs. R. Chaste and **P. Cassagne** (*in reply*): Among the different points outlined in the discussion, that of the loss of vertical resolution and misregistration inherent in the Secam system appears frequently in the remarks (Messrs. Cooper, Parker-Smith and Maurice) with that of the moiré patterns (Mr. Monteath). The experience with a large number of pictures does not show objectionable defects. If a ratio of luminance to chrominance bandwiths of three is considered suitable for the correct reproduction of pictures, the rise time of corresponding signals being in the inverse ratio, it may be shown that in the horizontal direction there is a misregistration of luminance and chrominance of at least one luminance picture element. In the Secam system the same colour signal is displayed on the lines n and $n + 2$, i.e. approximately on line $n + 1$, instead of its true position, line n. The misregistration is about the same.

Another important point, mentioned by Messrs. Parker-Smith, Jesty and Carnt, concerns the performance in the case of black-and-white pictures or eventual failure of chrominance signal. With amplitude modulation it was necessary, in order to follow conveniently the 'constant luminance' principle, to provide compensation of the d.c. component of the mean sub-carrier (not drawn in Fig. 4 of the paper) given by the d.c. component or the peak value of the burst. This has been realized without any supplementary valve in some receivers and operates correctly. Chrominance gain, level balance and white balance then act as in the N.T.S.C. system. However, this is unnecessary with frequency modulation, because, for the central frequency of the sub-carrier transmitted for the white, as in the absence of sub-carrier, the output of the discriminator is zero.

Messrs. Jesty and Townsend query the price of the delay line. The figure of £6 given in the paper can be considered as a retail price because it is that which could be obtained with a production of only 500 per month. It could be reduced by mass production and so would become cheaper than the six valves and the associated circuits of the N.T.S.C. receiver compared by Mr. Townsend.

The figures in Table 1 must not be taken as absolute values, but are given as relative thresholds for the visibility of interferences at different frequencies in the video spectrum for the amplitude-modulation case. In the frequency-modulation case the increase in protection is about 10 dB. We can give Mr. Kilvington no results for triangular noise.

In Section 8.2 it was not proposed to relate the field and mains frequencies, but only to reduce the difference between the two. This is possible by modifying the field and sub-carrier frequencies, which are always interrelated, and keeping the line frequency within the limits given for suitable registration of the delayed signal.

DISCUSSION ON

'THE APPLICATION OF TRANSISTORS TO LINE-COMMUNICATION EQUIPMENT'*

NORTH STAFFORDSHIRE SUB-CENTRE AT STONE, 22ND APRIL, 1960

Mr. K. A. MacKenzie: Reference has been made to the reliability and extremely long life of transistors. Modern practice is to employ printed-circuit techniques and package construction and to break down the circuit into a number of similar units which plug into the main assembly; the philosophy behind this is that, if a failure occurs, the faulty unit or package can be quickly located and replaced. This certainly simplifies maintenance and reduces the time any equipment is out of action, and to a certain extent it also permits first-line maintenance to be carried out by semi-skilled staff. Since the life expectancy of transistors and associated components when operated with reasonable design tolerances on all parameters is measured in terms of decades, is it correct to introduce possible unreliability in the form of plug-and-socket connections and additional terminations which may in themselves cause more failures? The justifications would appear to be rapid replacement if a fault does occur, provided that this is not on a socket or termination thereto, and economy of manufacture arising from standardized packages and assemblies. Will the authors comment on this?

Messrs. H. T. Prior, D. J. R. Chapman and **A. A. M. Whitehead** (*in reply*): Plug-and-socket connections can be made much more reliable than components such as resistors and capacitors, provided that cost, size and contact pressure are not important limiting factors. Where the total insertion or withdrawal force is great enough (and high contact pressure greatly assists reliability), extra cost and bulk may be added by insertion or withdrawal mechanisms or contact-locking mechanisms. Thus it is reasonable to provide high-reliability plugs and sockets where their size and cost are not too great compared with those of the units which they serve. The high life-expectancy of modern components is of great assistance in this respect, since it makes practicable the inclusion of hundreds of components in one plug-in assembly.

Units of this order of size are sometimes much larger than the basic packages from which the circuit can be constructed. This is particularly marked in digital equipment, where the basic packages might contain 10–100 components. Such packages may be plugged in or connected by soldered joints or solderless wrapped joints. The choice between these methods depends largely on equipment construction and maintenance techniques. We consider that soldered or wrapped joints are preferable, excepting where simple first-line maintenance is of overriding importance or where it is excessively difficult to make the connections accessible.

* Prior, H. T., Chapman, D. J. R., and Whitehead, A. A. M.: Paper No. 2722 R, September, 1958 (see 106 B, p. 279).

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A NEW H_{10}-TO-H_{20} MODE TRANSDUCER

By C. C. EAGLESFIELD, M.A., Associate Member, Y. KLINGER, Ph.D., and L. SOLYMAR.

(*The paper was first received* 24*th October*, 1959, *in revised form* 21*st March, and in final form* 3*rd June*, 1960.)

SUMMARY

Using a general method of mode-transducer design previously described, a simple H_{10}-to-H_{20} mode transducer was constructed. The reflection coefficient is calculated, and it is shown that for a transducer a few wavelengths in length the reflections are insignificant. The power transferred to the spurious modes was measured by the resonance method on two experimental models, 2 and 8 guide-wavelengths long. The longer transducer transforms 92% of the power into the required mode. The power in unwanted modes can be greatly reduced by the use of a simple absorptive mode filter.

(1) INTRODUCTION

The interest in mode transducers has been revived recently owing to new developments in microwave techniques.[1] The launching of the H_{01} mode in an overmoded circular waveguide is one of the topics of current interest.

The H_{01} mode of a circular waveguide can be excited from the H_{10} mode of a rectangular waveguide in two distinct steps:[1] first via a transducer to the H_{20} mode of a rectangular waveguide, then via a second transducer to the H_{01} mode of the circular waveguide. In the present paper a transducer of the first type is described. This is one of the simplest conceivable mode transducers: it can be easily designed with the aid of Reference 2, it can be easily constructed, and the simple geometry of its cross-section permits also some analytical calculations.

As shown in Reference 2, if a transducer is sufficiently long, any purity requirements can be satisfied. Since there are practical limitations to the length of an actual transducer, the power in the spurious modes generated by a short transducer needs to be studied.

The design method is briefly outlined in Section 2, whilst the calculation of the reflection coefficient can be found in Section 3. The power transferred to spurious modes was measured on two experimental models by the resonance method,[3] which is explained in Section 4. A way of eliminating the spurious modes is given in Section 5.

(2) OUTLINE OF THE DESIGN METHOD

The general design of mode transducers[2] is carried out with the aid of a suitably chosen eigenfunction which gradually varies between the eigenfunctions of the modes in the waveguides to be connected.

In the present case the eigenfunctions of the H_{10} mode in waveguide A and the H_{20} mode in waveguide B (Fig. 1) are identical. Thus the obvious choice for the eigenfunction of the transducer is

$$\psi = C \cos \frac{\pi x}{a} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The cross-section of the transducer has to be designed in such a way that the boundary condition $d\psi/dn = 0$ is satisfied at each

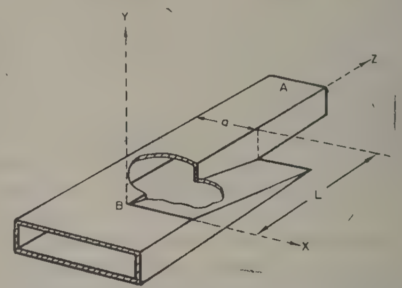

Fig. 1.—Mode transducer transforming the H_{20} mode of waveguide B into the H_{10} mode of waveguide A.

cross-section. This requirement is met in the transducer shown in Fig. 1, where at the point $z = 0$, half the broad face is made to taper down to meet the opposite wall in a sharp wedge, at the point $z = L$. The other half is left undisturbed and continues as waveguide A.

For simplicity, the variation of the height of the tapered section is chosen to be linear, but it must be borne in mind that an optimum choice of this curve might result in a considerable reduction in the amount of power in the spurious modes.

(3) CALCULATION OF THE REFLECTION COEFFICIENT

The amplitudes of the spurious modes can be calculated in principle with the aid of the formulae in Reference 4. However, as the eigenfunctions of the spurious modes are not known analytically and the numerical solution would require considerable work, there are no means at the moment of predicting theoretically the performance of this mode transducer.

The reflections, admittedly, are of much smaller importance, but since in the present case they are calculable it seems worth while to include the results.

The reflection coefficient in a general non-uniform waveguide is given by[4]

$$R = \int_0^L \left[S^- - \frac{1}{2} \frac{d(\log_\varepsilon K)}{dz} \right] \exp\left(-j2\int_0^z \beta dz\right) dz \quad . \quad (2)$$

where β = Phase-change coefficient.
K = Wave impedance.
S^- = Backward coupling coefficient, which for an H-wave can be expressed as follows:

$$S \quad -\tfrac{1}{2} \oint_{C(z)} \tan\theta \left(\frac{\partial\psi}{\partial s}\right)^2 ds \quad . \quad . \quad . \quad . \quad (3)$$

where $C(z)$ = The boundary.
ds = An element of the boundary.
θ = Angle between normal to $C(z)$ and normal to transducer.
ψ = The eigenfunction at a certain cross-section which satisfies the normalization condition

$$\int |\nabla\psi|^2 dS \quad 1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

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In the present case both the wave impedance and the propagation coefficient are constant, which greatly simplifies the expression for the reflection coefficient. The value of C in eqn. (1) can be calculated from eqn. (4), while $\tan\theta$ is given by $b'(z)$, the height of the tapered section. Substituting these into eqn. (2) and performing the mathematical operations,

$$R = \tfrac{1}{2}\int_0^L \frac{b'(z)}{b(0)+b(z)} \exp(-j2\beta z)dz \quad . \quad . \quad . \quad (5)$$

Assuming the linear variation in $b(z)$ and integrating, we get for the absolute value of the reflection coefficient

$$|R| = \tfrac{1}{2}\{[\mathrm{Ci}(4\beta L) - \mathrm{Ci}(2\beta L)]^2 + [\mathrm{Si}(4\beta L) - \mathrm{Si}(2\beta L)]^2\}^{1/2} \quad (6)$$

where
$$\mathrm{Ci}(t) = -\int_t^\infty \frac{\cos x}{x}dx \quad . \quad . \quad . \quad . \quad . \quad (7)$$

and
$$\mathrm{Si}(t) = \int_0^t \frac{\sin x}{x}dx \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

If βL is large enough, as it would be in practical applications, the asymptotic forms of the Si and Ci functions can be used,[5] leading to the simple result

$$R = \frac{1}{8\beta L}\sqrt{(5 - 4\cos 2\beta L)} \quad . \quad . \quad . \quad . \quad (9)$$

The maximum value of the reflection coefficient is

$$R_{max} \simeq 0{\cdot}06\frac{\lambda}{L} \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

which is sufficiently low for any practical purpose.

(4) MEASUREMENT OF THE SPURIOUS MODES

The transducer described produces spurious modes in waveguide B. The spurious modes capable of propagation in this waveguide are restricted to the H_{10} mode at low frequencies; at higher frequencies the H_{30} mode can also propagate.

The amount of power in the spurious modes generated in waveguide B was measured on two experimental models. Their lengths were chosen to be 2 and 8 guide wavelengths respectively.

The first transducer was deliberately made very short to enable us to study the case where the theoretical predictions are less valid. The $8\lambda_g$ transducer was expected to give reasonably low mode conversion.

The spurious modes were measured at a wavelength of approximately 9 mm, where the main mode propagating in the double guide was the required H_{20} and the only possible spurious mode was the H_{10} mode. The H_{30} mode was beyond cut-off and could not propagate.

The measurements were made by the resonance method. A good short-circuit introduced in the double guide could be positioned accurately by means of a micrometer. When the v.s.w.r. is measured in the single guide as a function of the position of this short-circuit, it is found that at regular intervals absorption lines, or resonances, appear. These are due to the spurious mode resonating between the transducer and the short-circuit when the latter is an appropriate distance away. The resonances are approximately half a guide wavelength apart (spurious mode), but not exactly. The magnitude of the reflection coefficient varies from one resonance to another in a regular manner. This can be understood easily with the help of Fig. 2, where the transducer is shown as coupling energy to the main mode and the suprious mode in series with each other. Clearly the spurious-mode line will resonate every time l is a multiple

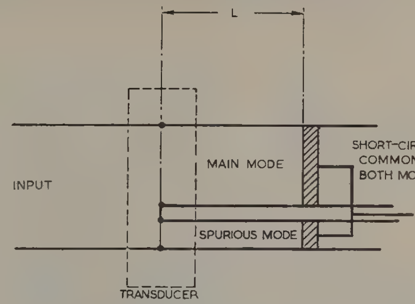

Fig. 2.—Schematic representation of the main mode and the spurious mode in the short-circuited section.

of $\frac{1}{2}\lambda_g$, extracting energy from the system and giving rise to resonance absorption at the input.

However, the impedance that the main mode presents at the junction also depends on the position of the piston, and whenever the main-mode line presents an open-circuit at the junction the reflection coefficient is unity. Thus, when a spurious-mode resonance is at or near coincidence with a main-mode open-circuit, the appropriate resonance will be missing, or at any rate very small.

These coincidences or near coincidences will occur once every half beat-wavelength. Thus at regular intervals of half a beat wavelength the resonances are very small or altogether missing. In between these points, the magnitude of the reflection coefficient at resonance varies from one to another; the envelope is shown in Fig. 3. Curves (a), (b) and (c) correspond to the low, critical

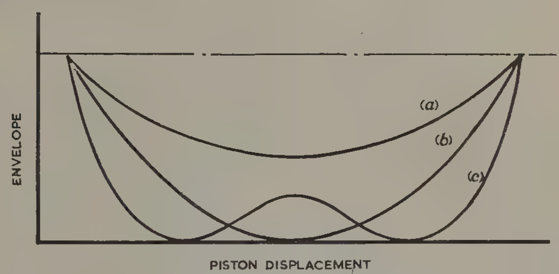

Fig. 3.—Envelope of the minima of the reflection coefficients as a function of piston movement.

(a) Low coupling.
(b) Critical coupling.
(c) High coupling.

and overcoupled cases, respectively, according as the ratio of loss to coupling into the spurious mode is larger than, equal to, or smaller than unity.

If the coupling is very much larger than the loss, resonances are not always obtained, but the presence of the spurious mode is revealed by a rapid change in phase of the reflection coefficient as the piston is moved through the resonance position.

In this case, by introducing sufficient loss into the system by a resistive card mounted on the piston, the resonances are obtained again, and their envelope corresponds to one of the curves shown in Fig. 3.

The resonance found equidistant from two 'missing' ones half a beat wavelength apart is measured accurately. From its shape it is possible to calculate the fraction of power coupled into the spurious mode. This fraction is given by[2]

$$|\Gamma|^2 = \frac{\pi l}{\lambda_g}(1 \mp |\Gamma_{min}|)\frac{1 - |\Gamma_0|^2}{|\Gamma_0|^2 - |\Gamma_{min}|^2} \quad . \quad . \quad . \quad . \quad (11)$$

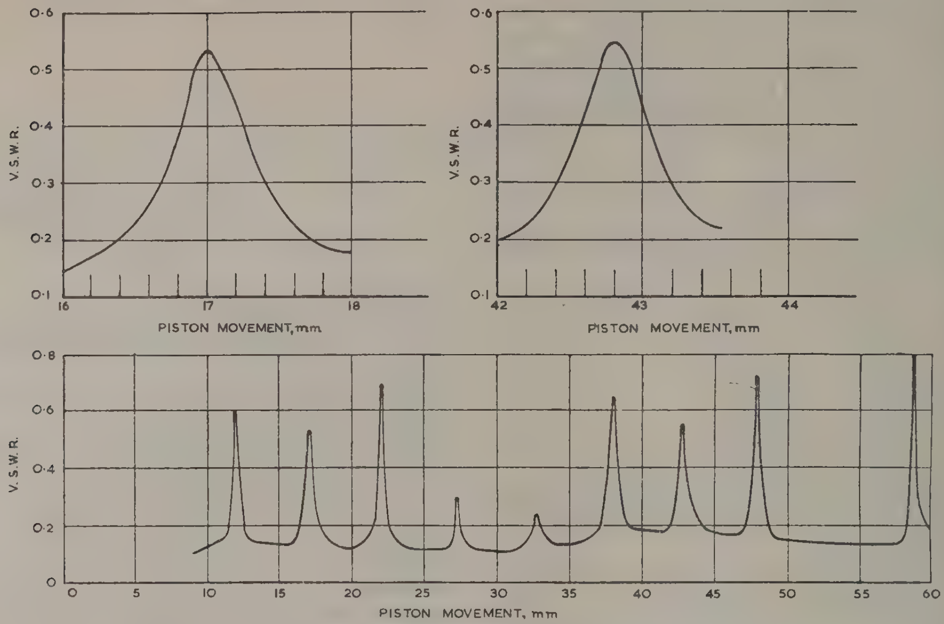

Fig. 4.—Relation between v.s.w.r. and piston movement for $2\lambda_g$ transducer.

where $|\Gamma|^2$ = Fraction of power in spurious mode.
$|\Gamma|_{min}$ = Magnitude of reflection coefficient at resonance.
$|\Gamma_0|$ = Magnitude of reflection coefficient when the piston is a distance l from the resonance position.
λ_g = Guide wavelength of spurious mode.

The sign in the first pair of brackets depends on the ratio of loss to coupling as shown in Fig. 3. For case (*a*) the sign is negative; for case (*c*), positive; in case (*b*), Γ_{min} is zero.

Fig. 4 illustrates the results of measurements on the $2\lambda_g$ transducer at $\lambda = 9{\cdot}75$ mm. Here the v.s.w.r. is shown as a function of piston position. In this case, the coupling was very much larger than the loss and it was necessary to introduce a resistive card on the front face of the piston. The curve illustrated was obtained in this way.

It can be seen that the general pattern of the resonances corresponds to the over-coupled case [curve (*c*) of Fig. 3]. The resonances are small or missing near piston positions 7, 30 and 53 mm. The resonances corresponding to eqn. (11) are found at about 17 and 43 mm.

To test the accuracy of the measurement, the resonance at 43 mm was analysed in detail and the coupling coefficient was calculated from the width of the resonance curve at different values of the v.s.w.r. The results are given in Table 1.

Table 1

V.S.W.R.	Γ_0	l	$\|\Gamma\|^2$
		mm	% power
0·54	0·299	—	—
0·50	0·333	0·107	27·2
0·45	0·379	0·17	26·9
0·40	0·429	0·23	27·0
0·35	0·481	0·30	27·6
0·30	0·538	0·40	30·0
0·25	0·600	0·52	32·0

$\lambda = 9{\cdot}75$ mm $\lambda_g = 13{\cdot}4$ mm
Size of waveguide A, 0·280 × 0·140 in.

The calculated value of the coupling is reasonably constant until the lower part of the resonance curve is approached, when the approximations used in deriving eqn. (11) no longer apply.

(5) ELIMINATION OF THE SPURIOUS MODES

The power in the spurious modes, as can be seen from Table 1, is about 27%. Although this is excessive for any practical use, the basic design considerations still seem to be valid. Even a transducer of this extremely short length shows a considerable discrimination in favour of the H_{20} mode. For the $8\lambda_g$ transducer the power in the spurious mode was found to be 8%.

The permitted amount of spurious mode is generally much less. However, in the present case we can get rid of it by the use of a simple mode filter. When an absorbing vane was put in the middle of waveguide B, no resonances were found, which suggests that the remaining impurity is less than about 0·3%. Thus the device is producing a practically pure H_{20} mode for the price of 8% loss of power. This is insignificant in a practical case, as most other devices show larger insertion loss. Furthermore, if for any reason the insertion loss must be kept low, the power in spurious modes could be reduced to a low value by making the transducer longer.

(6) ACKNOWLEDGMENT

The authors are grateful to Standard Telecommunication Laboratories Ltd. for permission to publish the paper.

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NON-REFLECTING WAVEGUIDE TAPERS

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SUMMARY

Much work has already been done towards establishing techniques for tapering the cross-section of waveguides without introducing significant reflections when waves are propagated along them. In the present paper the problem is approached from the aspect of maintaining a completely undisturbed field pattern outside the guide whatever its cross-section, so that the wave impedance remains constant. Examples are given illustrating how this result may, in principle, be achieved both for the hollow metal guides of rectangular and circular cross-section and for single-wire transmission lines. On the basis proposed it is shown that the surface impedance of the guide must always have a resistive component which has nothing to do with losses but is dependent solely upon the angle of the taper and represents energy crossing the interface from the field on one side to the field on the other.

(1) INTRODUCTION

When an electromagnetic wave is propagated along a boundary between two different media there is always a part of the field on each side of the interface and its distribution characterizes the wave mode or modes concerned. Both media store, and in general also dissipate, some of the energy of the wave. If the loss in one medium is high compared with that of the other, the mechanism for the support of the wave demands a flow of energy across the interface, giving rise to a resistive component of the surface impedance which will be positive or negative according to whether the net transfer of energy is respectively into or out of the surface. Moreover, for any given field distribution outside the surface there must be a corresponding reactive, as well as resistive, component to the surface impedance. Thus, focusing attention on a particular wave mode in a homogeneous medium, it is in principle always possible to erect a boundary surface at any point within the field without disturbing the wave, provided only that the surface is given the appropriate impedance. More or less of the field constituting the wave mode can thereby be completed within the surface, and this may be done without altering in any way the distribution of the field remaining outside the surface. The precise form taken by the field within the surface is immaterial and in most cases it will have a modified distribution; but so long as the conditions of continuity at the interface are satisfied, there will be no disturbance outside the surface. In general, therefore, a change in the position of the wall of a waveguide can be made if accompanied by an appropriate change of surface impedance. The proposition is a perfectly general one applying to all guided-wave modes which can exist separately when the wall of the guide has a finite impedance.[1] It is well known, for example, that the rate of decay of the field outside a single-wire transmission line supporting an axial cylindrical surface wave depends upon the radius of the guide and its surface reactance.[2] The larger this reactance and the smaller the size of the guide, the more rapidly does the field outside it die away, and one might therefore expect that a constant decay factor could be achieved by suitably adjusting the surface reactance for different radii, thus enabling the wave to propagate along a tapered guide without disturbance. Moreover, the counterpart of this arrangement, represented by the hollow tube supporting a waveguide mode within it, should clearly be susceptible to the same kind of treatment.

In a recent paper[3] the author called attention to methods of shifting the cut-off frequency of various tubular waveguide modes by appropriate changes of surface impedance. This technique can be interpreted as one which provides for more or less of the field of the wave to be accommodated inside the wall of the guide, which in turn can therefore be thought of as a cylindrical surface placed coaxially with the wave and having a surface impedance corresponding to that required for undisturbed support of the field within it. Precisely the same kind of interpretation is applicable to the surface wave, except that the whole arrangement is then turned inside out.

It seems therefore reasonable to expect that waveguide tapers for a wide variety of applications can be designed on the basis of an appropriate change of surface impedance accompanying the change of cross-sectional dimensions of the guide, so that no significant discontinuity arises in the wave impedance. Waves having circular symmetry are obviously particularly suited to this kind of treatment, but modes supported by rectangular guides can also be dealt with in much the same way.

The paper discusses a number of examples which appear in principle to meet the requirements, special attention being given to the axial cylindrical surface wave, the H_{01} and E_{01} modes in a circular guide and the H_{01} mode in a rectangular guide.

Since there is always a part of the field on each side of the supporting surface, so also is the total energy of the wave similarly distributed. For a lossless guide of uniform cross-section the energy in the field on each side of the interface remains constant. On the other hand, a tapered supporting surface arranged so as not to disturb the field pattern outside it requires a transfer of energy across the boundary as the wave progresses, in accordance with the change in the field accommodated on each side of the boundary, and this is necessary irrespective of any losses in the associated media. Thus, for the tapered guide there must always be a resistive component to the surface impedance. This quantity is dependent on the angle of the taper and is inherent in the mechanism envisaged. Such a surface resistance does not imply an energy loss but merely a movement of energy across the interface from one part of the field to another. In designing a tapered guide capable of supporting a given wave mode in accordance with the principles discussed, it is essential to make suitable provision for the reactive component of the required surface impedance; but when this has been accomplished, the corresponding resistive component, as a feature of the taper alone, should take care of itself. In general there will, of course, be a small supplementary surface resistance arising from the losses in the surrounding media, and due allowance should be made for this in matching the fields at the interface. In all cases the taper can apparently take any form desired, provided that the boundary conditions essential to the undisturbed support of the required wave mode

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

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are maintained, but for simplicity in the examples given a linear taper of small angle has been considered.

It should be emphasized that the practicability of these proposals has not yet been checked. Towards that end a programme of experimental work coupled with more detailed analysis is being undertaken.

(2) TAPERED CYLINDRICAL GUIDES FOR SURFACE WAVES

In terms of cylindrical co-ordinates, as shown in Fig. 1, an axial surface wave travelling in the $+x$ direction has field components in medium 1 (supposed to be air) outside the guide given by

$$E_{x1} = C_1\varepsilon^{-\gamma x}\mathrm{H}_0^{(1)}(h_1 r) \quad . \quad . \quad . \quad . \quad . \quad (1)$$

$$H_{\theta 1} = C_1\frac{j\omega\epsilon_0}{h_1}\varepsilon^{-\gamma x}\mathrm{H}_1^{(1)}(h_1 r) \quad . \quad . \quad . \quad . \quad (2)$$

$$E_{r1} = C_1\frac{\gamma}{h_1}\varepsilon^{-\gamma x}\mathrm{H}_1^{(1)}(h_1 r) \quad . \quad . \quad . \quad . \quad . \quad (3)$$

where
$$h_1^2 = -u_1^2 = \gamma^2 + K_1^2 \quad . \quad . \quad . \quad . \quad . \quad (4)$$

$$K_1^2 = \omega^2\mu_0\epsilon_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

$$\gamma = \alpha + j\beta \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

$$u_1 = a_1 - jb_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

and the usual sinusoidal time variation represented by $\varepsilon^{j\omega t}$ is assumed.

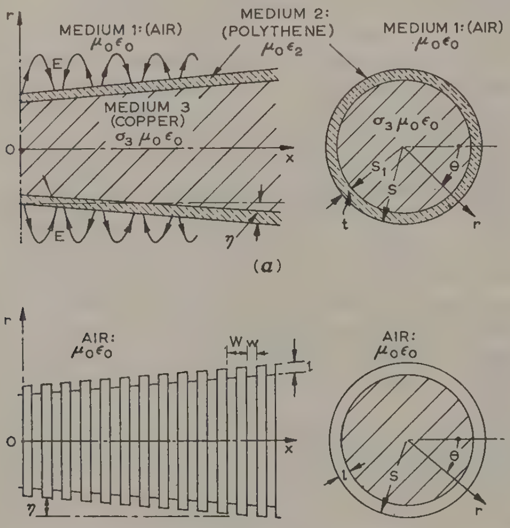


Fig. 1.—Tapered waveguides for axial cylindrical surface wave.
(a) Dielectric-coated guide.
(b) Corrugated guide.

If the supporting surface of radius s is tapered, as defined by the angle, η, it makes with the axis of the guide,

$$\eta = \frac{ds}{dx} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

and to avoid any disturbance of the wave outside the surface we must ensure continuity of the tangential components of the electric and magnetic fields across the boundary. Thus at $r =$ we require to match the tangential electric field

$$E_{t1} = E_{x1}\cos\eta + E_{r1}\sin\eta \quad . \quad . \quad . \quad . \quad (9$$

and the tangential magnetic field, $H_{\theta 1}$, with the correspondin field components in medium 2.

For a loss-free guide the required conditions can be met b progressively increasing the surface reactance as the radius increased, and this may in principle be carried out with the hel of either a dielectric-coated or a corrugated conductor acting a the supporting structure. In the first case [Fig. 1(a)] the thicl ness of the dielectric coating must increase with s, and in th second [Fig. 1(b)] the depth of the transverse corrugations mu change in a similar way. The surface impedance of the tapere guide with regard to the surface wave whose field componen are given by equations (1), (2), (3) and (9) is

$$Z_s = R_s + jX_s = \left[\frac{E_{t1}}{H_{\theta 1}}\right]_{r=s} = \frac{h_1\cos\eta}{j\omega\epsilon_0}\frac{\mathrm{H}_0^{(1)}(h_1 s)}{\mathrm{H}_1^{(1)}(h_1 s)} + \frac{\gamma\sin\eta}{j\omega\epsilon_0} \quad . \quad (1$$

For negligible losses in the surface, eqn. (6) reduces to $\gamma = j$ and using eqns. (4), (5) and (7) we find

$$u_1^2 = \beta^2 - \omega^2\mu_0\varepsilon_0 \quad . \quad . \quad . \quad . \quad . \quad (1$$

This is a 'slow' wave with $\beta^2 > \omega^2\mu_0\varepsilon_0$, so that $u_1 = a_1$ is pure real. We can therefore pick out the resistive and reactive con ponents of Z_s given by eqn. (10), i.e.

$$R_s = \frac{\beta\sin\eta}{\omega\epsilon_0} \quad . \quad . \quad . \quad . \quad . \quad . \quad (1$$

$$X_s = \frac{a_1\cos\eta}{j\omega\epsilon_0}\frac{\mathrm{H}_0^{(1)}(ja_1 s)}{\mathrm{H}_1^{(1)}(ja_1 s)} = \frac{a_1\cos\eta}{\omega\epsilon_0}\frac{\mathrm{K}_0(a_1 s)}{\mathrm{K}_1(a_1 s)} \quad . \quad (1$$

Thus, although we have assumed no losses, there is a surfac resistance R_s, providing for the power crossing the interface, as result of the taper, from the storage field on one side to th storage field on the other. If, as shown in Fig. 1, the radius o the guide is increasing as the wave progresses from left to righ then $\eta = ds/dx$ is positive and power flows into the surface wit R_s positive. On the other hand, for a wave in the opposi direction supported by this guide, η is negative and so conse quently is R_s. For small angles η we have $\cos\eta \simeq 1$ an therefore the surface reactance X_s from eqn. (13) become practically the same as that required by a purely cylindric guide of the same radius. With this in mind we can now procee to determine the conditions appropriate to a dielectric-coate metal guide with a small taper.

(2.1) Dielectric-Coated Metal Guide

For the purpose of calculating the surface reactance of dielectric-coated metal guide having negligible losses and onl a small angle of taper, we shall consider it to be of purel cylindrical form. Actually we must have slightly different taper on the inside and the outside of the dielectric coating, sinc $t = (s - s_1)$ changes a little along the length of the guid although the effect of the angle of the taper as distinct fron variations in the thickness of the dielectric is relatively very smal

Thus for medium 2 in Fig. 1(a) we can write

$$E_{x2} = C_2'\varepsilon^{-\gamma x}\mathrm{J}_0(h_2 r) + C_2''\varepsilon^{-\gamma x}\mathrm{Y}_0(h_2 r) \quad . \quad . \quad (1$$

$$H_{\theta 2} = C_2'\frac{j\omega\epsilon_2}{h_2}\varepsilon^{-\gamma x}\mathrm{J}_1(h_2 r) + C_2''\frac{j\omega\epsilon_2}{h_2}\varepsilon^{-\gamma x}\mathrm{Y}_1(h_2 r) \quad . \quad (1$$

$$E_{r2} = C_2'\frac{\gamma}{h_2}\varepsilon^{-\gamma x}\mathrm{J}_1(h_2 r) + C_2''\frac{\gamma}{h_2}\varepsilon^{-\gamma x}\mathrm{Y}_1(h_2 r) \quad . \quad (1$$

here $$h_2^2 = -u_2^2 = \gamma^2 + K_2^2 \quad . \quad . \quad . \quad . \quad (17)$$

$$K_2^2 = \omega^2\mu_0\epsilon_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (18)$$

$$u_2 = a_2 - jb_2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (19)$$

nd for medium 3 we have

$$E_{x3} = C_3\varepsilon^{-\gamma x}J_0(h_3r) \quad . \quad . \quad . \quad . \quad . \quad (20)$$

$$H_{\theta 3} = C_3\frac{\sigma_3}{h_3}\varepsilon^{-\gamma x}J_1(h_3r) \quad . \quad . \quad . \quad . \quad (21)$$

$$E_{r3} = C_3\frac{\gamma}{h_3}\varepsilon^{-\gamma x}J_1(h_3r) \quad . \quad . \quad . \quad . \quad (22)$$

here $$h_3^2 = -u_3^2 = \gamma^2 + K_3^2 \quad . \quad . \quad . \quad . \quad (23)$$

$_3 \gg \omega\epsilon_3$, so that

$$K_3^2 = -j\omega\mu_0\sigma_3 \quad . \quad . \quad . \quad . \quad . \quad (24)$$

$$u_3 = a_3 - jb_3 \quad . \quad . \quad . \quad . \quad . \quad . \quad (25)$$

To meet the requirements we must therefore have at the outside ırface of the dielectric

$$jX_s = \left[\frac{E_{x2}}{H_{\theta 2}}\right]_{r=s} = \frac{h_2}{j\omega\epsilon_2}\,\frac{J_0(h_2s) + \frac{C_2''}{C_2'}Y_0(h_2s)}{J_1(h_2s) + \frac{C_2''}{C_2'}Y_1(h_2s)} \quad . \quad (26)$$

ence matching the tangential components of E and H at the ıterface where $r = s_1$ and remembering that since $|h_3s_1| \gg 1$ we ın put

$$\frac{J_0(h_3s_1)}{J_1(h_3s_1)} \simeq -j \quad . \quad . \quad . \quad . \quad . \quad . \quad (27)$$

/e then find $$\frac{C_2''}{C_2'} = \frac{\frac{\omega\epsilon_2h_3}{\sigma_3h_2}J_1(h_2s_1) - J_0(h_2s_1)}{Y_0(h_2s_1) - \frac{\omega\epsilon_2h_3}{\sigma_3h_2}Y_1(h_2s_1)} \quad . \quad . \quad . \quad (28)$$

rom eqns. (26) and (28) we can therefore calculate the value X_s for any given thickness of dielectric coating over a metal ɔd of specified diameter. Subject only to the limitation that ıe angle η must be small, we can then proceed to match the ılue of X_s derived in this way to the corresponding value given y eqn. (13), which is necessary for the support of the surface ave.

As an example of such a calculation we will take the case of a ɔlythene-coated copper conductor of circular cross-section ıpporting an axial cylindrical surface wave in air at a frequency of 10^{10} c/s. We will suppose that we require to maintain a constant decay coefficient $a_1 = 65\,\mathrm{m}^{-1}$ for the field outside the guide, that the angle of the taper has a constant value of $\eta = 1°$, that the relative permittivity of the dielectric coating, ϵ_r, is 2·2 and that the conductivity of the copper, σ_3, is 6×10^7 mhos/m.

Starting with a 38 s.w.g. copper wire coated with polythene 3 mils thick, making the overall radius of the guide 0·0152 cm and tapering this up to 9·24 cm over a length of 5·28 m, we find, neglecting losses, a constant $R_s = 6{\cdot}9$ ohms with values of X_s and $t = (s - s_1)$ as shown in Table 1.

It is interesting to observe that, when $s = \infty$, corresponding to a flat surface, we require for the same field decay $a_1 = 65\,\mathrm{m}^{-1}$ the value $X_s = 117$ ohms, which is obtained with $t = 0{\cdot}272$ cm.

The manner in which X_s is required to change with guide radius is shown in Fig. 2.

Table 1

Radius of guide, s	Surface reactance, X_s	Dielectric-coated guide: Radius of conductor, s_1	Dielectric-coated guide: Thickness of dielectric coating, t	Corrugated guide: Depth of slot, l
cm	ohms	cm	cm	cm
0·0152	5·5	0·0076	0·0076	—
0·053	13·4	0·028	0·0254	—
0·077	18·7	0·0414	0·0356	—
0·308	42·7	0·218	0·09	0·106
0·77	65·6	0·625	0·145	0·16
1·54	82	1·355	0·185	0·196
4·62	101	4·382	0·238	0·235
9·24	109	8·99	0·25	0·25

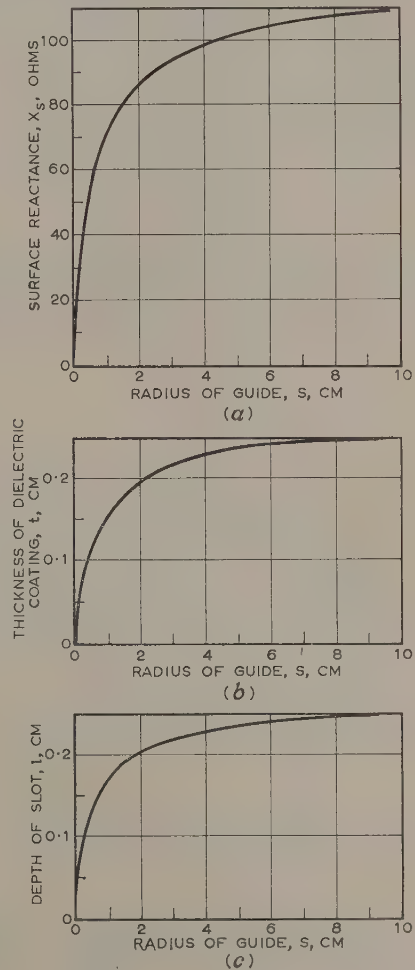


Fig. 2.—Surface waveguide for $f = 10^{10}$ c/s, $a_1 = 65\,\mathrm{m}^{-1}$ and $\eta = 1°$.
(a) Surface reactance, X_s.
(b) Thickness, t, of dielectric.
(c) Slot depth, l.

(2.2) Corrugated Metal Guide

In a corrugated metal guide transverse corrugations, each supporting a radial TEM wave, are appropriate in so far as they provide for the required surface reactance. There is, however, another important consideration which must not be overlooked.

Since power is progressively transferred across the supporting surface as the wave travels forward, there must be a flow of energy along the interface and through the medium on each side of it. There is no difficulty here with the dielectric-coated smooth metal surface, considered in Section 2.1, but when separate transverse slots are introduced, the manner in which provision is made for the essential coupling between them is not at once obvious. It is suggested that the coupling mechanism rests upon the fact that the effective supporting surface for the wave is some little distance from the mouth of the slots and that energy passes from one slot to the next in the intermediate region. The TEM waves in the slots cannot themselves match up to the surface wave, and there must be a transition field which includes a travelling wave just outside the openings to the slots. This view of the situation is apparently also applicable, although in a more limited way, to the well-known purely cylindrical guide with transverse corrugations. Deliberate coupling holes might be provided between adjacent slots when large angles of taper are involved.

If the circumferential slots of width w and pitch W have a depth l which is small compared with the radius, s, of the guide [Fig. 1(b)], the surface reactance is given by

$$X_s = \frac{w}{W}\sqrt{\frac{\mu_0}{\epsilon_0}}\tan\frac{2\pi l}{\lambda} \quad . \quad . \quad . \quad . \quad . \quad (29)$$

With $w/W = \frac{1}{2}$, calculations have been made for the guide sizes considered in Section 2.1 and under conditions for which eqn. (29) is sufficiently accurate. These results are given in Table 1 and in Fig. 2.

and neglecting losses

$$K^2 = \omega^2\mu_0\epsilon_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad .$$

$$\gamma \doteq j\beta \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad .$$

The magnetic field tangential to the surface of the tapered gu at $r = s$ is

$$H_t = H_x \cos\eta + H_r \sin\eta \quad . \quad . \quad . \quad .$$

giving the surface impedance

$$Z_s = R_s + jX_s = \left[\frac{E_\theta}{H_t}\right]_{r=s} = -\frac{j\omega\mu}{h\cos\eta\dfrac{J_0(hs)}{J_1(hs)} + j\beta\sin\eta} \quad .$$

or

$$R_s = -\frac{\omega\mu P}{P^2 + Q^2} \quad . \quad . \quad . \quad . \quad . \quad .$$

and

$$X_s = -\frac{\omega\mu Q}{P^2 + Q^2} \quad . \quad . \quad . \quad . \quad . \quad .$$

where

$$Q = h\cos\eta\frac{J_0(hs)}{J_1(hs)} \quad . \quad . \quad . \quad . \quad . \quad .$$

and

$$P = \beta\sin\eta \quad . \quad . \quad . \quad . \quad . \quad . \quad .$$

As an example of the requirements to be met in a particular ca suppose $f = 3\cdot5 \times 10^{10}$ c/s, $\eta = 1°$ and choose a tube such t $Z_s = 0$ at $s = 1\cdot5$ cm, implying that under these conditi $J_1(hs) = 0$, $(hs) = 3\cdot831$, so that $h = 3\cdot831/0\cdot015 = 255$ a $\beta = 686$. To preserve the wave pattern within the tape guide we must keep $h = 255$ as a constant.

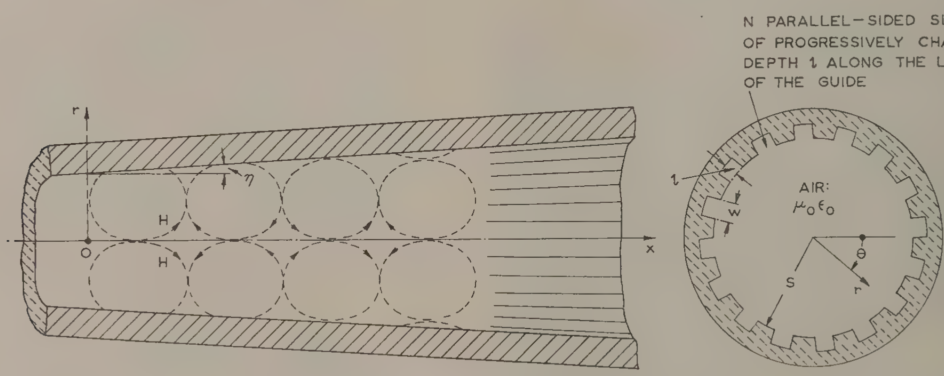


Fig. 3.—Tapered waveguide for circular H_{01} mode.

(3) TAPERED GUIDES FOR CYLINDRICAL WAVEGUIDE MODES

(3.1) H_{01} Wave Mode

In Fig. 3 the field components of the H_{01} mode supported inside a hollow guide of circular cross-section and propagated in the $+x$ direction are given by

$$H_x = C\varepsilon^{-\gamma x}J_0(hr) \quad . \quad . \quad . \quad . \quad . \quad . \quad (30)$$

$$H_r = C\frac{\gamma}{h}\varepsilon^{-\gamma x}J_1(hr) \quad . \quad . \quad . \quad . \quad . \quad . \quad (31)$$

$$E_\theta = -C\frac{j\omega\mu}{h}\varepsilon^{-\gamma x}J_1(hr) \quad . \quad . \quad . \quad . \quad . \quad (32)$$

where

$$h^2 = \gamma^2 + K^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (33)$$

Calculating from eqns. (38) and (39) the required resistive a reactive components of the surface impedance for values o ranging from $1\cdot5$ to $2\cdot75$ cm, we get the curves in Fig. 4. R_s negative in this case, because power must come out of the surfa as the wave progresses.

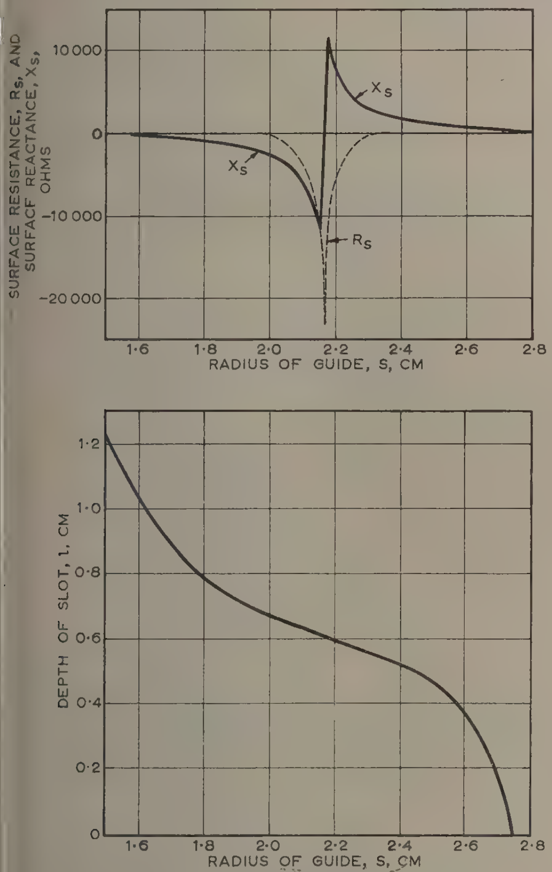
To satisfy the requirements in regard to X_s, the wall of t guide can conveniently be provided with longitudinal sl symmetrically disposed around the circumference, as shown Fig. 3. If there are N parallel-sided slots, each of width w a of progressively changing depth l as s varies along the length the guide, the average surface reactance presented to the wave a given cross-section is

$$X_s \simeq \frac{Nw\omega\mu}{2\pi sh}\tan(hl) \quad . \quad . \quad . \quad . \quad .$$

For a chosen distribution of slots we can match this value of

o the corresponding quantity calculated from eqn. (39), and or small angles of taper the appropriate depth of slot l at any oint along the guide can thus be determined.

In order to provide in the example given for a suitable variation f surface reactance, the slots are made about half a wavelength eep when $s = 1{\cdot}5$ cm and $X_s = 0$. The results of the calcula- on for $N = 16$ and $w = 0{\cdot}3$ cm are shown in Fig. 4, but it hould be borne in mind that, owing to the radial distribution f the field, these results tend to be only very approximate when the tube is small and the slots comparatively deep. For more accurate calculation in such circumstances the ratio of the circular functions in eqn. (42) should be replaced by the more precise Bessel functions.



g. 4.—Tubular waveguide with longitudinal corrugations supporting circular H_{01} mode.

$f = 3{\cdot}5 \times 10^{10}$ c/s, $h = 255$ and $\eta = 1°$

(3.2) $\mathbf{E_{01}}$ Wave Mode

For the $\mathbf{E}_{01}$ mode (Fig. 5) the field components are

$$E_x = C\varepsilon^{-\gamma x}J_0(hr) \quad . \quad . \quad . \quad . \quad . \quad . \quad (43)$$

$$E_r = C\frac{\gamma}{h}\varepsilon^{-\gamma x}J_1(hr) \quad . \quad . \quad . \quad . \quad . \quad . \quad (44)$$

$$H_\theta = C\frac{j\omega\epsilon}{h}\varepsilon^{-\gamma x}J_1(hr) \quad . \quad . \quad . \quad . \quad . \quad . \quad (45)$$

where
$$h^2 = \gamma^2 + K^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (46)$$

and for negligible losses

$$K^2 = \omega^2\mu_0\epsilon_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (47)$$

$$\gamma = j\beta \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (48)$$

The electric field tangential to the wall of the tapered guide is

$$E_t = E_x \cos\eta + E_r \sin\eta \quad . \quad . \quad . \quad . \quad . \quad (49)$$

and the surface impedance is

$$Z_s = R_s + jX_s = -\left[\frac{E_t}{H_\theta}\right]_{r=s} = \frac{jh\cos\eta}{\omega\epsilon_0}\frac{J_0(hs)}{J_1(hs)} - \frac{\beta\sin\eta}{\omega\epsilon_0} \ . \ (50)$$

giving
$$R_s = -\frac{\beta}{\omega\epsilon_0}\sin\eta \quad . \quad . \quad . \quad . \quad . \quad . \quad (51)$$

and
$$X_s = \frac{h\cos\eta}{\omega\epsilon_0}\frac{J_0(hs)}{J_1(hs)} \quad . \quad . \quad . \quad . \quad . \quad . \quad (52)$$

If we again take a tapered tube with $\eta = 1°$ operating at $f = 3{\cdot}5 \times 10^{10}$ c/s and having $X_s = 0$ at $s = 1{\cdot}5$ cm, it follows that $h = 160{\cdot}3$ with $\beta = 714$ and Fig. 6 gives the required R_s and X_s values at different radii. Circumferential slots of width w, pitch W and depth l [as used on the outside of the surface waveguide discussed in Section (2.2)] are apparently applicable here, and provided that $l \ll s$, eqn. (29) can be applied to give an average surface reactance matching the corresponding value derived from eqn. (52). Fig. 6 shows the progressive change in the depth of the slots required for this taper.

The problem of coupling between the adjacent slots again arises here and can, it seems, be met only if the mechanism discussed in Section 2.2 is valid.

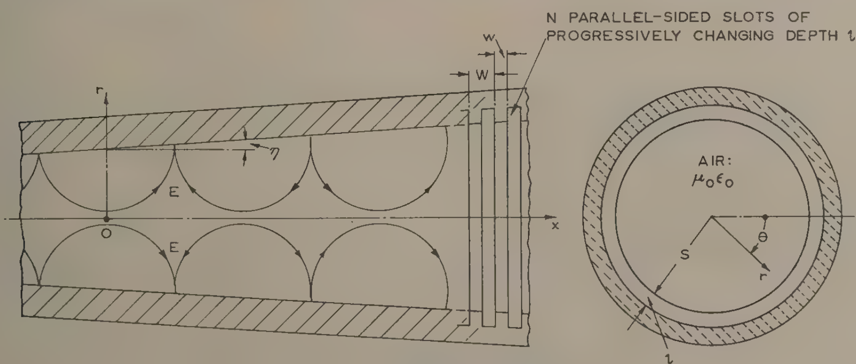


Fig. 5.—Tapered waveguide for circular E_{01} mode.

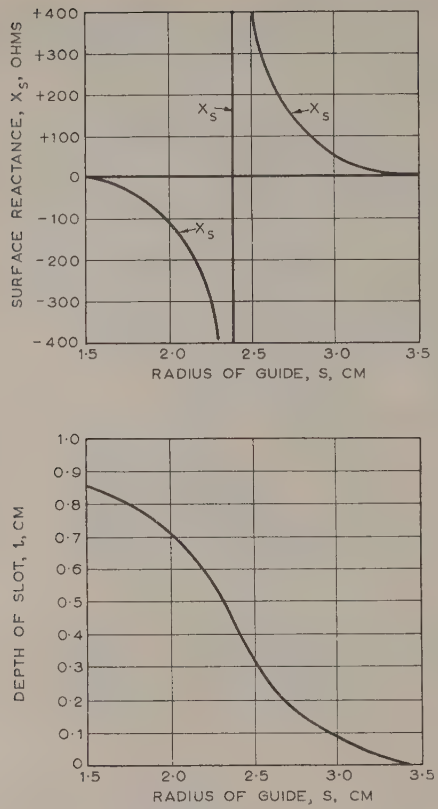


Fig. 6.—Tubular waveguide with transverse corrugations supporting E_{01} mode.

$= 3{\cdot}5 \times 10^{10}$ c/s, $h = 160{\cdot}3$, $\eta = 1°$, $w/W = \frac{1}{2}$, $R_s = -6{\cdot}43$ ohms and is constant.

(4) TAPERED GUIDE OF RECTANGULAR CROSS-SECTION SUPPORTING THE DOMINANT MODE

In this case it is clear that the narrow face of the guide (E-plane) can present to the dominant H_{01} mode a finite wall impedance without introducing any other modes. This is not so with the broad face of the guide, and consequently to apply the kind of taper we have been discussing the change of cross-sectional dimension must be in the H-plane.

Referring to Fig. 7, in which the origin of the Cartesian co-ordinates is conveniently taken at the centre of one of the broad faces of the tapered guide, we can write the field co
ponents as

$$E_y = C\varepsilon^{-\gamma x}\cos(hz)$$

$$H_x = C\frac{jh}{\omega\mu}\varepsilon^{-\gamma x}\sin(hz)$$

$$H_z = C\frac{\gamma}{j\omega\mu}\varepsilon^{-\gamma x}\cos(hz)$$

where $$h^2 = \gamma^2 + K^2 = (\pi/b_0)^2$$

and for negligible losses

$$K^2 = \omega^2\mu_0\epsilon_0$$

$$\gamma = j\beta$$

The magnetic field tangential to the wall of the guide is

$$H_t = H_x\cos\eta + H_z\sin\eta$$

and the corresponding surface impedance of the narrow face
the guide is

$$Z_s = R_s + jX_s = -\left[\frac{E_y}{H_t}\right]_{z=\pm b/2}$$

$$= -\frac{\omega\mu}{\beta\sin\eta + j\dfrac{\pi}{b_0}\cos\eta\tan\left(\dfrac{\pi b}{2b_0}\right)}$$

b_0 being the value of b for which $Z_s = 0$;

or $$R_s = -\frac{\omega\mu P}{P^2 + Q^2}$$

and $$X_s = \frac{\omega\mu Q}{P^2 + Q^2}$$

where $$Q = \pi/b_0\cos\eta\tan\frac{\pi b}{2b_0}$$

and $$P = \beta\sin\eta$$

It will be observed that R_s is negative for the forward-travell
wave with power coming out of the tapered wall of the guide

Taking the case of a rectangular guide tapered in the H-pla
so that $\eta = 1°$, operated at $f = 10^{10}$ c/s and having zero surf
impedance when $b_0 = 2{\cdot}54$ cm, we find R_s and X_s values
given in Fig. 8. Over a limited range of change of b the narr
faces of the guide could be dielectric coated to satisfy the requ
ments, or for a larger range use can be made of longitudinal s
(Fig. 7). If there are N slots each of width w and dept
uniformly distributed over each narrow face of the guide, th

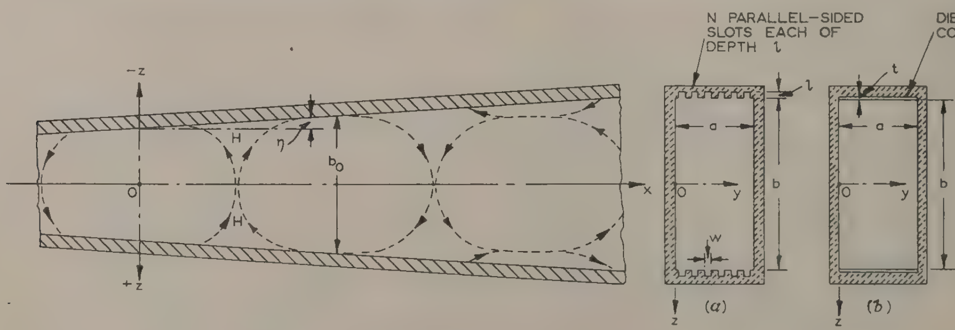


Fig. 7.—Tapered waveguide for rectangular H_{01} mode.

(a) Narrow face longitudinally corrugated.
(b) Narrow face dielectric coated.

eglecting the small effect of the taper, these slots present an verage surface reactance

$$X_s = \frac{Nw\omega\mu}{ah} \tan (hl) \quad . \quad . \quad . \quad . \quad . \quad (65)$$

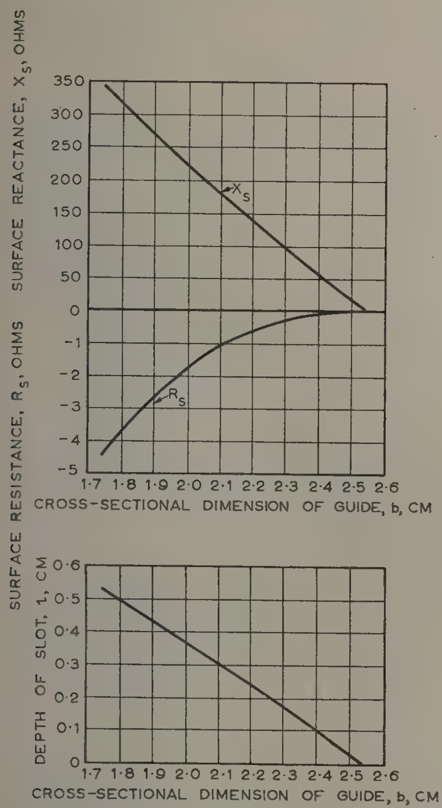


Fig. 8.—Rectangular waveguide with longitudinal corrugations supporting H_{01} mode.

$f = 10^{10}$ c/s, $\beta = 169$, $\eta = 1°$ and $b_0 = 2 \cdot 54$ cm.

Assuming $N = 6$, $a = 1 \cdot 27$ cm and $w = 0 \cdot 15$ cm, we find that slots of depth shown in Fig. 8 satisfy the requirements.

(5) CONCLUSIONS

A technique has been proposed, and is tentatively supported by a number of preliminary calculations, for tapering the cross-sectional dimensions of waveguides without introducing any significant change in the wave impedance. This development appears to offer a wide variety of applications, but it must be borne in mind that no experimental work on the subject has so far been carried out and that a more precise analysis is necessary to check in detail some of the calculations made.

In the examples given a uniform taper has been assumed, but there seems nothing to prevent other contours from being employed if so desired, provided always that the appropriate surface impedance is presented to the wave.

It will be observed that all the E-modes, in contrast to the H-modes, require a relatively small and constant value of R_s. This makes the problem to which reference is made in Section 2.2 of providing for the necessary coupling between adjacent transverse slots less exacting.

(6) ACKNOWLEDGMENTS

The author is indebted to Dr. J. Brown, Mr. J. B. Davies and Mr. P. H. Hargrave for helpful discussions on this subject.

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A NEW CAVITY-RESONATOR METHOD FOR MEASURING PERMITTIVITY

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SUMMARY

The difficulties of the usual cavity-resonator methods of measuring permittivity are discussed, and details are given of a new method which eliminates most of these at the expense of increasing the computation required to interpret the results. The theory of the new method is described in full, and typical examples of experimental results obtained by its use are given.

LIST OF PRINCIPAL SYMBOLS

a = Radius of cavity.
b = Radius of rod.
d = Length of empty cavity section.
l = Length of dielectric rod.
$m = b/a$.
K = Turns ratio of transformer.
$p = 3\cdot 832/a = (k_0^2 - \beta_1^2)^{1/2}$.
$p_1 = (\epsilon_r k_0^2 - \beta_2^2)^{1/2}$.
$p_2 = (k_0^2 - \beta_2^2)^{1/2}$.
$x = p_1 b$.
$y = p_2 a$.
k_0 = Free-space plane-wave phase-change coefficient, $2\pi/\lambda_0$.
$F(y) = J_0(my)\,Y_1(y) - Y_0(my)J_1(y)$.
$G(y) = J_1(my)Y_1(y) - Y_1(my)J_1(y)$.
Q, Q' = Cavity Q-factors.

$$R = \sin^2(\beta_1 d) + \left(\frac{\beta_1}{\beta_2}\right)^2 \cos^2(\beta_1 d).$$

$$S = \frac{\beta_2 a^2 p^2}{4\beta_1 x^2 y^2}\left\{4x^2 - \pi^2 my(x^2 - m^2y^2)[myF(y) - 2G(y)]F(y)\right\}$$

$$T = \frac{\pi^2 a^2 m^2 p^2}{4\beta_1 x^2}[m^2y^2F^2(y) - 2myF(y)G(y) + x^2G^2(y)]$$

β_1 = Phase-change coefficient of H_{01} mode in empty cavity.
β_2 = Phase-change coefficient of H_{01} mode in cavity with dielectric rod.
α_1 = Attenuation coefficient of H_{01} mode in empty cavity.
α_2 = Attenuation coefficient of H_{01} mode in cavity with dielectric rod.
ϵ_r = Relative permittivity of rod.
δ = Loss angle of dielectric rod.
J, Y = Bessel functions of first and second kinds.

(1) INTRODUCTION

Several methods of using cavity resonators to measure the dielectric properties of materials at microwave frequencies have been developed. Three such methods are widely used in different situations and are as follows:

(*a*) Disc samples in an H_{01n} cavity.[1, 2, 3]
(*b*) An axial rod in an E_{010} cavity.[1, 3]
(*c*) Small samples in any suitable cavity.[4, 5]

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

Dr. Sinha and Dr. Brown are at University College, London. The paper forms part of a thesis submitted by one of the authors (J.K.S.) for the degree of Doctor of Philosophy at London University.

The first method uses the cavity mode with the highest Q-fac and is therefore particularly suited for measurements on lo loss materials. The second method is useful at longer wa lengths, since the size of the cavity required to support the E mode is appreciably smaller than that for an H_{01n} mode. T third method needs much smaller material samples than other two and is used for relatively lossy materials.

Each of these methods has its own difficulties. Methods and (*b*) require that the dielectric sample should be an accur fit with respect to one of the cavity dimensions. The disc method (*a*) must have the same diameter as the cavity, while rod in method (*b*) must have the same length as the cavity. T provision of samples with the required dimensions is relativ straightforward for materials such as plastics, which are ea machined, but it can be very tedious when ceramic mater are being investigated. Further, there appears to be no sim test by which the measurements can be checked to ensure t a small gap between the cavity walls and the dielectric is introducing errors.

In methods (*a*) and (*c*), the measurements can be made a selected frequency, the cavity being tuned to resonate at t frequency. In method (*b*), however, the cavity resonant quency cannot be altered and resonance must be obtained adjusting the oscillator frequency. It is not always conveni to do this.

The equations from which the relative permittivities are culated, if method (*c*) is used, are based on a perturbation of cavity fields and are therefore only approximate. This le to the possibility of errors in the relative permittivities wh are deduced from the measurements. Such errors can eliminated by making measurements on a series of samples different sizes,[6, 7] but this considerably increases the time nee to obtain the results.

The object of the work described in the paper was to deve a cavity method which would be free from the difficulties lined above. The major requirements are as follows:

(*a*) The samples used should not require to have any dimens fixed to specified values.
(*b*) The measurement should be made at a fixed freque resonance being obtained by tuning the cavity.
(*c*) Contacts between the dielectric and the cavity walls sh be avoided, so that problems associated with the presence of s air gaps do not arise.
(*d*) The theory from which the dielectric properties are culated should be exact, to avoid the uncertainties associated the perturbation method.

It is obviously also desirable that a single cavity shoul capable of giving results for materials with a wide rang permittivity and loss tangent, and if possible, for magr materials, such as ferrites.

(2) OUTLINE OF THE METHOD

The method which has been developed uses a cylindrical which is inserted in progressive steps along the axis of a cy drical cavity operating in the H_{01n} mode. The cross-sectio the cavity with the rod in position is shown in Fig. 1. The is held by a collet connected to a micrometer spindle, and

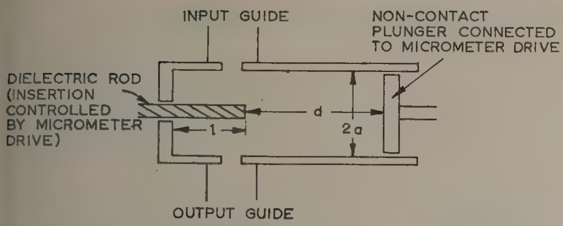

Fig. 1.—General arrangement of the cavity and dielectric sample.

e moved along the axis, thus varying the rod insertion l. For ich rod position, the cavity is tuned to resonance at the fre- uency selected for the measurements by adjusting the non- ontact plunger. The hole through which the rod enters the ivity can be changed in diameter by inserting an appropriate ush. The diameter of the hole is selected to ensure that the rcular waveguide which it forms is well below cut-off at the perating frequency, even when filled by the dielectric being easured. From the readings obtained, the total length of the ivity at resonance $(d + l)$ is plotted against the rod insertion, l, e resulting curve being identical in form to those obtained by e well-known Feenberg or Weissfloch method[8, 9] which is sed to examine the effect of discontinuities in waveguides. The avelength of the H_{01} mode in the section of the cavity con- ining the rod can easily be determined from this curve, and ence the permittivity of the rod material can be calculated.

In principle, the loss tangent can be calculated from a corre- onding curve in which the cavity Q-factor is plotted against e rod insertion. The delineation of such a curve in sufficient etail is, however, a very tedious matter, and a simpler procedure as been used in which only two values of Q-factor need be easured, these being taken for two positions of the rod for hich l differs by half the wavelength of the mode in the cavity ction containing the rod. It will be shown in the next Section at this procedure eliminates any errors arising from reductions the Q-factor caused by the cavity end walls or the coupling les.

The cavity used in the experimental work operated in the $_{013}$ mode and has been described in an earlier paper.[10] Input d output couplings to rectangular waveguides are provided circular holes at diametrically opposite positions on the vity side wall. The fixed end-plate of the original cavity was placed by the assembly, which controls the rod insertion.

(3) THEORY OF THE METHOD

(3.1) Permittivity Measurements

The relation between l and $(d + l)$, corresponding to resonance a specified angular free-space wavelength λ_0, can be derived, for any high-Q-factor cavity, by ignoring losses. An equi- lent circuit which represents the cavity is shown in Fig. 2. e transmission line 1 corresponds to the empty section of the vity and has the phase-change coefficient, β_1, of the H_{01} ode in an empty circular waveguide of radius a, equal to that the cavity. Hence

$$\beta_1^2 = (2\pi/\lambda_0)^2 - p^2 \qquad (1)$$

here

$$p = 3{\cdot}832/a \qquad (2)$$

ansmission line 2 represents the cavity section containing the electric rod, and its phase-change coefficient, β_2, is that of the $_{01}$ mode in a circular guide of radius a, containing a dielectric d of relative permittivity ϵ_r and radius b. The coefficient β_2 is a complicated function of a, b, λ_0 and ϵ_r. The first three can be measured, so that ϵ_r can be calculated if β_2 is also measured.

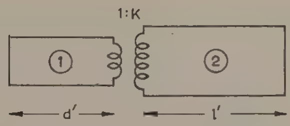

Fig. 2.—Equivalent circuit for the cavity.

Transmission lines 1 and 2 correspond, respectively, to the H_{01} modes in the empty cavity and the cavity with a coaxial dielectric rod.

The characteristic impedances of both lines in Fig. 2 can be taken as unity, the impedance discontinuity at the junction between the empty and filled sections being accounted for by the ideal transformer of turns ratio K. The lengths of the lines in the equivalent circuit are taken as d', l', and these may differ from the corresponding cavity distances, d, l, for two reasons. First, there are small phase shifts at the discontinuity between the empty and filled sections, and secondly, the end walls of the cavity may not be perfect short-circuits. Since losses can be ignored, the cavity end walls can be represented as lumped reactances, which, in turn, can be expressed as lengths of the lines 1 and 2. Hence

$$d' = d + d_1 \qquad (3)$$

$$l' = l + l_1 \qquad (4)$$

where d_1, l_1 are the additional line lengths to represent the above two effects. Although d_1, l_1 will depend on frequency, they will be constant for a set of measurements at one fixed frequency.

The condition that the circuit of Fig. 2 should resonate at the free-space wavelength λ_0 is

$$\tan(\beta_2 l') = -K^2 \tan(\beta_1 d') \qquad (5)$$

This is an equation identical to that obtained in the Weissfloch method.[8, 9] Eqn. (5), coupled with the Weissfloch theory, can therefore be used to predict the form of the dependence between l' and d'. In the present application, it is most convenient to plot $(d + l)$ against l. The curve of $(d' + l')$ against l' has the shape shown in Fig. 3, and since d', l' differ from d, l by constant

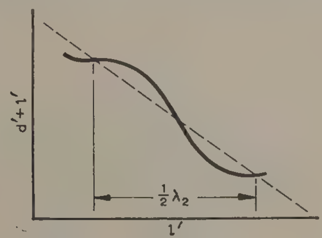

Fig. 3.—General shape of the curve showing the dependence of $(d' + l')$ on l'.

amounts, the curve of $(d + l)$ against l will have the same shape, but be displaced by l_1 along the horizontal axis and by $(d_1 + l_1)$ along the vertical axis. The mean slope of the curve, i.e. the slope of the dotted line, is $-(\beta_2 - \beta_1)/\beta_1$. This slope can be measured from the experimental curve; β_1 can be calculated using eqn. (1) or can be measured in the empty cavity, so that β_2 is available from the measured slope.

The only remaining step is the calculation of ϵ_r from the measured value of β_2. The relationship between β_2, ϵ_r, a, b and λ_0 has been derived by Pincherle[11] and can be expressed by the following set of equations:

Let
$$k_0 = 2\pi/\lambda_0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$
$$m = b/a \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$
$$x^2 = (\epsilon_r k_0^2 - \beta_2^2)b^2 = p_1^2 b^2 \quad . \quad . \quad . \quad . \quad . \quad (8)$$
$$y^2 = (k_0^2 - \beta_2^2)a^2 = p_2^2 a^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

Then
$$\frac{J_1(x)}{xJ_0(x)} = \frac{J_1(my)Y_1(y) - J_1(y)Y_1(my)}{my[J_0(my)Y_1(y) - J_1(y)Y_0(my)]} \quad . \quad (10)$$

In the present problem, k_0, β_2, a and b are obtained directly from the measurements and ϵ_r is to be calculated. This may be done as follows:

(i) Calculate m and y from eqns. (7) and (9), respectively.
(ii) Substitute these values in the right-hand side of eqn. (10) and solve for x. There are an infinite number of possible roots, of which the smallest corresponds to the H_{01} mode and is therefore the one required.
(iii) Substitute the value of x, obtained from eqn. (2), in eqn. (8) and calculate ϵ_r.

The difficult step is the solution of eqn. (10) for x and this is facilitated by the preparation of numerical Tables of the functions in the equation. A limited number of calculations has been carried out for the values of k_0, a and b which were used in experimental work, and the results have been used to prepare the curves of Fig. 4. It would be a relatively straightforward matter to prepare a complete set of results appropriate to any values of the parameters by using a digital computer.

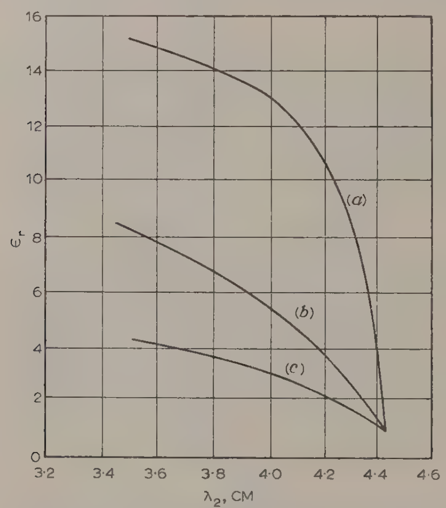

Fig. 4.—Dependence of the relative permittivity ϵ_r on the wavelength λ_2 of the H_{01} mode in the cavity section containing the dielectric rod.

The curves are calculated for a cavity radius of 2·82 cm and a free-space wavelength of 3·20 cm. The radius of the dielectric rod is
(a) 0·318 cm.
(b) 0·475 cm.
(c) 0·635 cm.

The success of this method requires that only one of the H_{0n} modes in the cavity section containing the dielectric should propagate. This places a restriction on the maximum value of ϵ_r for given values of λ_0, a and b, and this maximum can be calculated by imposing the condition that it corresponds to the cut-off of the H_{02} mode. Fig. 5 shows the restriction on the values of ϵ_r for the experimental conditions used.

Eqn. (10) is based on the assumption that y is real, and this implies that β_2 is less than k_0 [see eqn. (9)]. When the limit given by the results in Fig. 5 is approached, β_2 may become

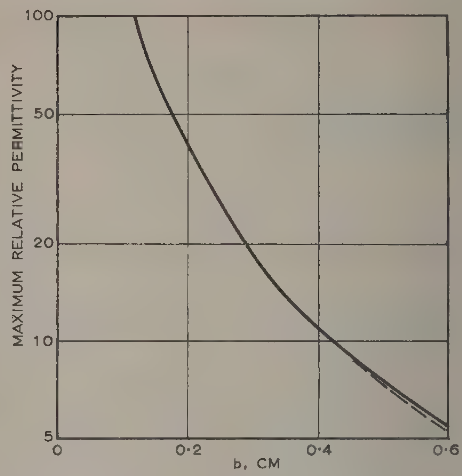

Fig. 5.—Maximum relative permittivity which can be measured f rod of given radius in a cavity of radius 2·82 cm at a free-sp wavelength of 3·20 cm.

——— H_{02} cut-off.
– – – – Condition that the wavelength of the H_{01} mode in the section conta the dielectric rod equals the free-space wavelength.

greater than k_0 and eqn. (10) requires modification. To av this and at the same time to ensure that the H_{02} mode is not near cut-off, a second restriction that β_2 should not exceed is imposed. If $\beta_2 = k_0$, then $y = 0$ and eqn. (10) reduces to

$$\frac{J_1(x')}{x'J_0(x')} = -\frac{1 - m^2}{2m^2} \quad . \quad . \quad . \quad .$$

where
$$x' = (\epsilon_r' - 1)^{1/2} k_0 b \quad . \quad . \quad . \quad .$$

ϵ_r' being the value of ϵ_r which gives $\beta_2 = k_0$. These results easily be plotted in a universal form, as in Fig. 6. The co

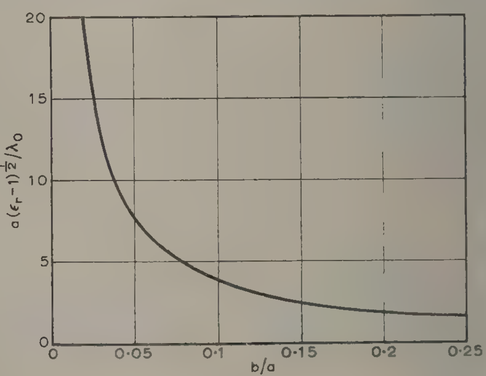

Fig. 6.—Universal curve showing the condition that the wavele of the H_{01} mode in the cavity section with the dielectric ro radius b should equal the free-space wavelength λ_0.

sponding restriction for the conditions used in the experime work is shown in Fig. 5, for comparison with the restric based on H_{02} cut-off.

(3.2) Loss-Tangent Measurements

The loss tangent of the material under test is determined in all cavity methods, by measurements of the Q-factor. major difficulty to be overcome is to distinguish betw

e various sources of loss which contribute to the Q-factor. the present problem, the possible sources of loss are as llows:

Conduction in the cavity side wall.
Conduction in the cavity end walls.
Loss in the material under test.
Radiation through the coupling holes to the input and output guides.

further complication is that the losses in the cavity side walls ay be larger than those predicted from a consideration of the $_{01}$ modes, because additional H_{0n} evanescent modes are cited at the discontinuity between the empty section and the ction containing the dielectric rod. The procedure which s been adopted is to make a pair of Q-factor measure-ents in which as many of the above factors as possible are kept nstant.

Consider the cavity arrangement shown in Fig. 1. The neral definition for the Q-factor is

$$\omega \times \frac{\text{Total stored energy in the cavity}}{\text{Power loss in the cavity}} \quad . \quad . \quad . \quad (13)$$

here ω = Resonant angular frequency.

This equation can be written as

$$Q = \frac{\omega(W_1 + W_2)}{P_1 + P_2 + P_3 + P_4 + P_5} \quad . \quad . \quad . \quad (14)$$

here

W_1 = Total stored energy in the empty section.
W_2 = Total stored energy in the section containing the rod.
P_1 = Power lost by the attenuation in the side walls of the H_{01} mode in the empty section.
P_2 = Power lost by the attenuation in the side walls and in the dielectric rod of the H_{01} mode in the section containing the rod.
P_3 = Power lost through the coupling holes.
P_4 = Power lost in the end walls.
P_5 = Power lost in the side walls because of the excitation of evanescent modes at the discontinuity.

Insert the rod by an extra half wavelength (measured in the ction containing the rod) at the same time decreasing the ngth of the empty section by a half wavelength (measured the empty section). Eqn. (5) shows that the cavity will still e resonant. Further, an examination of the field equations ows that the peak values of the standing-wave fields in the o sections of the cavity can be taken to be the same in this se as in the one considered above. A consequence of this is at the loss terms P_3, P_4 and P_5 will be identical in the two ses. The new Q-factor will therefore be

$$Q' = \frac{\omega[W_1' + W_2']}{P_1' + P_2' + P_3 + P_4 + P_5} \quad . \quad . \quad . \quad (15)$$

here W_1', W_2', P_1' and P_2' have the same significance as W_1, $_2$, P_1 and P_2, except that they apply to the increased rod sertion.

The terms P_3, P_4 and P_5 can therefore be eliminated from ns. (14) and (15), giving

$$P_1' + P_2' - P_1 - P_2 = \frac{\omega(W_1' + W_2')}{Q'} - \frac{\omega(W_1 + W_2)}{Q} \quad (16)$$

The various quantities in this equation can be expressed in rms of the cavity fields and the attenuation of the H_{01} modes the two sections of the cavity. The attenuation of the H_{01} ode in the section containing the dielectric rod can be expressed a form (Section 7.2) which depends on the loss tangent of the dielectric and on the attenuation of the H_{01} mode in the empty section of the cavity. The calculations required to obtain an equation for tan δ in terms of the Q-factors are given in Section 8 and lead to the result:

$$\epsilon_r RT \tan\delta = \frac{2\alpha_1\beta_2(S-R)}{k_0^2} + \frac{1}{Q'}\left[\left(R - \frac{\beta_2}{\beta_1}\right)S + (\epsilon_r - 1)RT\right]$$
$$+ \frac{1}{\pi}\left(\frac{1}{Q'} - \frac{1}{Q}\right)\left\{\beta_2(d + Rl)S + (\epsilon_r - 1)\beta_2 lRT\right.$$
$$\left. + \frac{1}{2}\left[\left(\frac{\beta_1}{\beta_2} - \frac{\beta_2}{\beta_1}\right)S + \frac{\beta_1}{\beta_2}(\epsilon_r - 1)T\right]\sin(2\beta_1 d)\right\} \quad (17)$$

where α_1 is the attenuation of the H_{01} mode in the empty section of the cavity, and the functions R, S and T are given by the equations:

$$R = \sin^2(\beta_1 d) + \left(\frac{\beta_1}{\beta_2}\right)^2 \cos^2(\beta_1 d) \quad . \quad . \quad . \quad . \quad . \quad . \quad (18)$$

$$S = \frac{\beta_2 a^2 p^2}{4\beta_1 x^2 y^2}\{(4x^2 - \pi^2 my(x^2 - m^2y^2)\mathrm{F}(y)[my\mathrm{F}(y) - 2\mathrm{G}(y)]\} \quad (19)$$

$$T = \frac{\pi^2 m^2 a^2 p^2}{4\beta_1 x^2}[m^2y^2\mathrm{F}^2(y) - 2my\mathrm{F}(y)\mathrm{G}(y) + x^2\mathrm{G}^2(y)] \quad (20)$$

where

$$\mathrm{F}(y) = \mathrm{J}_0(my)\mathrm{Y}_1(y) - \mathrm{J}_1(y)\mathrm{Y}_0(my) \quad . \quad . \quad (21)$$

$$\mathrm{G}(y) = \mathrm{J}_1(my)\mathrm{Y}_1(y) - \mathrm{J}_1(y)\mathrm{Y}_1(my) \quad . \quad . \quad (22)$$

All the quantities involved in eqn. (17), except tan δ, can be measured or computed from measured values. Tan δ can thus be calculated.

(4) EXPERIMENTAL RESULTS

Measurements on a variety of dielectrics have been made by the method described in the previous Section, and those discussed here are selected to illustrate the typical performance. Full details of these and other measurements are contained elsewhere.[12] The cavity used has been previously described by El-Ibiary and Brown[10] and was modified by replacing the fixed end-plate by the rod-insertion mechanism discussed in Section 2. The radius of the cavity is 2·82 cm, and all measurements were made at the fixed frequency of 9·375 Gc/s to minimize the computational work involved in interpreting the results. It should be emphasized that all the ancillary microwave equipment was of ordinary commercial standard and that the object of the experimental work was to verify the principles of the method. A considerable increase in accuracy could be achieved by using an oscillator whose frequency and amplitude are stabilized and by improving the techniques used to measure the Q-factors.

(4.1) Measurements on Solid Dielectrics

Rods of relative permittivities ranging from 2·5 (polystyrene) to 90 (titanium dioxide) have been measured, and consistent results in agreement with published data have been obtained.

Typical measured curves showing the variation of the total cavity length with the rod insertion are given in Fig. 7 and show that the experimental points lie on a curve of the predicted shape. The relative permittivity is deduced from a curve of this shape, as discussed in Section 3.1.

Q-factors were deduced from measurements of the insertion attenuation of the cavity at resonance. This gives the loaded Q-factor, but since losses by radiation through the coupling

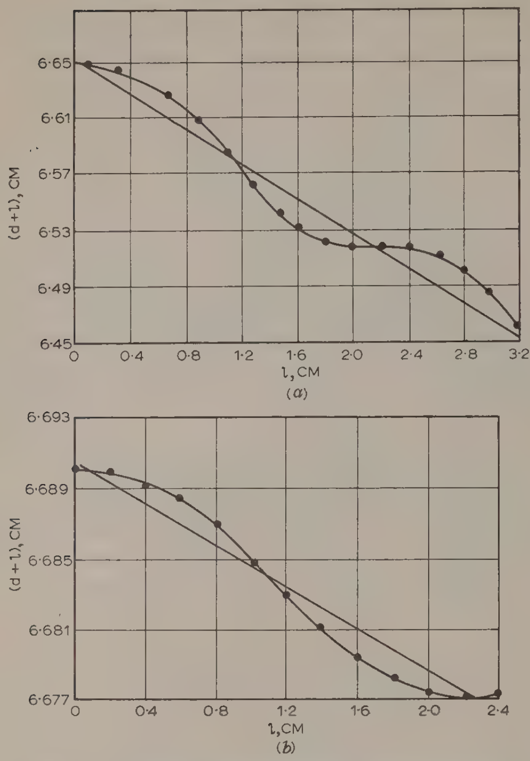

Fig. 7.—Typical measured curves of $(d + l)$ against l.

(a) Polystyrene rod of radius 0·635 cm.
(b) Titanium dioxide rod of radius 0·105 cm.

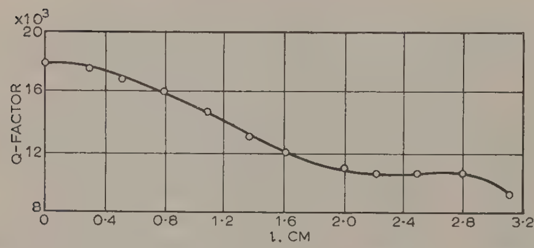

Fig. 8.—Dependence of Q-factor on rod insertion for a polystyrene rod of radius 0·635 cm.

holes are eliminated by the procedure described in Section 3.2, it is unnecessary to correct these to unloaded values. A typical curve of the variation of Q-factor with rod insertion is shown in Fig. 8. From this curve, a series of pairs of values separated by half a wavelength was taken, and the attenuation in the cavity section containing the dielectric was calculated from each pair. The values so obtained were consistent to within $\pm 3\%$, this being the estimated accuracy of the Q-factor measurement. The loss tangent is calculated from this attenuation coefficient and that of the empty cavity, which is deduced from the Q-factors of H_{011} and H_{012} resonances.

The effect of the sample diameter on loss-tangent measurements was investigated by making measurements on three Perspex rods, prepared from one piece of material. The results are given in Table 1 and are again consistent within the limits of the accuracy of the Q-factor measurements.

The oscillatory character of the curve showing Q against l is

Table 1

RESULTS ON PERSPEX SAMPLES MEASURED AT 9·375 Gc/s

Sample diameter	Measured values	
	Relative permittivity	Loss tangent
in		
0·250	2·62	0·0092
0·375	2·66	0·0094
0·500	2·65	0·0095

due to the mismatch arising at the junction between the t cavity sections. This mismatch can be eliminated by adding quarter-wave matching section to the end of the rod, a it is then found that a linear relation between simple fu tions of Q and l can be deduced. Full details of the theo of this method and of results obtained by its use are giv elsewhere.[12]

(4.2) Measurements on Liquids

A number of methods by which liquids can be measured cavity resonators have been described.[13,15] These involve container for the liquid, and a similar technique can be us with the present cavity. The liquid is contained in a cylindri bottle which is inserted in steps along the cavity axis, just as solid materials. The theory is rather complicated, since involves the determination of the characteristics of a wavegu with three materials; air, the container wall and the liqu Otherwise the method is essentially similar to that for so materials and can give quite reliable results.[12]

(4.3) Measurements on Ferrites

The complex permittivity and permeability of unmagneti ferrites may be determined by using this cavity system. T rod specimens of different diameters are used and measureme are made on each to determine the attenuation coefficient a the wavelength in the partially filled portion of the cavity. Fr these results, the constants of the material are evaluated. T evaluation of the constants in this method is based on exact theoretical solution, as in the method of Roberts a Srivastava.[16] The details of the theory and results obtained measurements with samples of low-loss ferrites are given e where.[12] This method has been found to give results co parable in accuracy with those obtained by other techniques.

(5) DISCUSSION

The advantages of the method described in the paper can summarized as follows:

(a) The relative permittivity is obtained from a series measurements instead of from single observations, as in previ cavity methods. The method is a precision one in the se discussed by Oliner and Altschuler.[17]

(b) A similar precision technique could, in principle, be appl to the determination of loss tangent, but the computatio labour is prohibitive. The accuracy of the method can, ho ever, be improved by taking a series of pairs of Q-factors a averaging the loss tangents so obtained. The possibility several measurements of loss tangent minimizes the likeliho that errors in the measurements will not be detected.

(c) The dielectric specimen does not require any dimension have a specified value. This simplifies the preparation ceramic specimens.

(*d*) The measurements can be made for rods of a reasonable range of diameters (see Table 1). It is thus possible to make measurements on a single rod in different cavities to cover a range of frequencies.

(*e*) As with any cavity method, the results are accurate only if the fields have the form assumed in the theoretical calculations. In particular, the excitation of unwanted propagating modes will seriously affect the accuracy. A major advantage of the method described is that the presence of unwanted modes is shown by a deviation of the experimental curve of cavity length against rod insertion from the expected shape. This has been confirmed experimentally by using a rod which was not circular and thus excited H_{11} and E_{11} modes.

The only major disadvantage is the computational work required in extracting the results. This can be materially reduced if measurements are made at selected fixed frequencies and detailed numerical charts are computed.

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(7) APPENDICES

(7.1) Properties of a Circular Guide with a Coaxial Dielectric Rod

(7.1.1) Field Components.

The behaviour of a circular guide with a coaxial dielectric rod has been studied by Pincherle,[11] and the results required in the present paper follow by a similar analysis. The notation used follows that in the main text. Within the dielectric rod, i.e. for $0 \leqslant r \leqslant b$, the field components for the H_{01} mode are

$$E_\theta = -\frac{j\omega\mu_0}{p_1}AJ_1(p_1r) \quad . \quad . \quad . \quad . \quad . \quad (23)$$

$$H_r = -Y_2E_\theta \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (24)$$

$$H_z = AJ_0(p_1r) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (25)$$

where
$$p_1^2 = \epsilon_r k_0^2 - \beta_2^2 : Y_2 = \beta_2/\omega\mu_0 \quad . \quad . \quad . \quad (26)$$

and the factor $\exp(j\omega t - j\beta_2 z)$ is omitted from all the components. Outside the rod, i.e. $b \leqslant r \leqslant a$,

$$E_\theta = -\frac{j\omega\mu_0}{p_2}[BJ_1(p_2r) + CY_1(p_2r)] = -\frac{j\omega\mu_0}{p_2}Z_1(p_2r) \quad (27)$$

$$H_r = -Y_2E_\theta \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (28)$$

$$H_2 = BJ_0(p_2r) + CY_0(p_2r) = Z_0(p_2r) \quad . \quad . \quad . \quad . \quad . \quad (29)$$

where
$$p_2^2 = k_0^2 - \beta_2^2 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (30)$$

The amplitude constants A, B and C must be selected to satisfy the boundary conditions at $r = a$ and $r = b$, and these lead to the relations

$$\frac{B}{Y_1(y)} = -\frac{C}{J_1(y)} = \frac{J_0(x)A}{J_0(my)Y_1(y) - J_1(y)Y_0(my)} \quad (31)$$

together with eqn. (10) from which the phase-change coefficient is obtained.

The quantities x, y and m are defined in eqns. (7)–(9).

(7.2) Attenuation

Attenuation arises from dielectric losses and the conductivity of the waveguide wall, and can be calculated by the usual method, assuming that the field patterns are not materially different from those for the loss-free case. The attenuation constant is given by

$$\alpha_2 = \frac{1}{2P}\frac{dP}{dz} \quad . \quad . \quad . \quad . \quad . \quad . \quad (32)$$

where P is the power flow through the guide cross-section and $\frac{dP}{dz}$ is the power loss per unit length of guide. From Poynting's theorem:

$$P = \tfrac{1}{2}\int_{cross\text{-}section} \mathscr{R}(E_\theta H_r^*)dA$$

$$= \pi Y_2 \int_0^a |E_\theta|^2 r dr$$

$$= \pi Y_2 \omega^2 \mu_0^2 \left[\frac{|A|^2}{p_1^2}\int_0^b r J_1^2(p_1 r)dr + \frac{1}{p_2^2}\int_b^a r Z_1^2(p_2 r)dr\right] \qquad (33)$$

The integrals are of the standard Lommel type and can be evaluated to give

$$P = \pi\omega\mu_0\beta_2\left\{\frac{|A|^2 b^2}{2p_1^2}\left[J_1^2(x) - \frac{2}{x}J_0(x)J_1(x) + J_0^2(x)\right]\right.$$

$$\left. + \frac{a^2}{2p_2^2}Z_0^2(y) - \frac{b^2}{2p_2^2}\left[Z_1^2(my) - \frac{2}{my}Z_1(my)Z_0(my) + Z_0^2(my)\right]\right\} \qquad (34)$$

in which the result that $Z_1(y) = 0$ has been used.

Also

$$\frac{Z_1(my)}{p_2} = \frac{AJ_1(x)}{p_1} \text{ and } Z_0(my) = AJ_0(x) \qquad (35)$$

because of the boundary conditions at $r = b$. Hence

$$P = \tfrac{1}{2}\pi\omega\mu_0\beta_2\left\{|A|^2 b^2\left(\frac{1}{p_1^2} - \frac{1}{p_2^2}\right)\left[J_0^2(x) - \frac{2J_0(x)J_1(x)}{x}\right] + \frac{a^2 Z_0^2(y)}{p_2^2}\right\} \qquad (36)$$

The loss per unit length due to the conductivity of the walls is

$$(dP/dz)_c = \tfrac{1}{2}R_m \int_0^{2\pi} |H_2|^2_{r=a} a d\theta$$

$$= \pi a R_m Z_0^2(y) \qquad (37)$$

where R_m = Surface resistance of the metal.

The loss per unit length in the dielectric rod is

$$(dP/dz)_d = \tfrac{1}{2}\omega\epsilon_r\epsilon_0 \tan\delta \int_0^{2\pi}\int_0^b |E_\theta|^2 r dr d\theta$$

$$= \frac{\pi a^3 \mu_0^2 \epsilon_r \epsilon_0 |A|^2 b^2 \tan\delta}{2p_1^2}\left[J_1^2(x) - \frac{2}{x}J_0(x)J_1(x) + J_0^2(x)\right]$$

where $\tan\delta$ is the loss tangent of the dielectric. Substituti[ng] these results in eqn. (32) gives

$$\alpha_2 = \frac{aR_m Z_0^2(y) + \dfrac{\omega^3\mu_0^2\epsilon_r\epsilon_0|A|^2 b^2\tan\delta}{2p_1^2}\left[J_1^2(x) - \dfrac{2}{x}J_0(x)J_1(x) + J_0^2(x)\right]}{\omega\mu_0\beta_2\left\{|A|^2 b^2\left(\dfrac{1}{p_1^2} - \dfrac{1}{p_2^2}\right)\left[J_0^2(x) - \dfrac{2}{x}J_0(x)J_1(x)\right] + \dfrac{a^2}{p_2^2}Z_0^2(y)\right\}}$$

The surface resistance R_m can be expressed in terms of the attenuation coefficient of the H_{01} mode in the empty circu[lar] guide. The equation for the empty guide corresponding [to] eqn. (39) is

$$\alpha_1 = \frac{p^2 R_m}{\omega\mu_0\beta_1 a}$$

where β_1 is the phase-change coefficient in the empty guide

and
$$p^2 = k_0^2 - \beta_1^2$$

Further,
$$Z_0(y) = BJ_0(y) + CY_0(y)$$

$$= \frac{AJ_0(x)[J_0(y)Y_1(y) - J_1(y)Y_0(y)]}{J_0(my)Y_1(y) - J_1(y)Y_0(my)}$$

$$= -\frac{2AJ_0(x)}{\pi y F(y)}$$

where
$$F(y) = J_0(my)Y_1(y) - Y_0(my)J_1(y)$$

Then

$$\alpha_2 = \frac{\dfrac{4\beta_1 a^2 J_0^2(x)\alpha_1}{\pi^2 p^2 y^2} + \dfrac{\omega^2\mu_0\epsilon_0\epsilon_r b^2 F^2(y)}{2p_1^2}\left[J_1^2(x) - \dfrac{2}{x}J_0(x)J_1(x) + J_0^2(x)\right]\tan\delta}{\beta_2\left\{b^2F^2(y)\left(\dfrac{1}{p_1^2} - \dfrac{1}{p_2^2}\right)\left[J_0^2(x) - \dfrac{2}{x}J_0(x)J_1(x)\right] + \dfrac{4a^2J_0^2(x)}{\pi^2 p_2^2 y^2}\right\}}$$

This rather formidable expression can be computed most eas[ily] if the Bessel functions of argument x are replaced by usi[ng] eqn. (10). Also pa is equal to 3·832, the first root of $J_1(x)$.

Let
$$G(y) = J_1(my)Y_1(y) - J_1(y)Y_1(my)$$

Then
$$\frac{J_1(x)}{xJ_0(x)} = \frac{G(y)}{myF(y)}$$

and

$$\alpha_2 = \frac{8\beta_1\alpha_1(xy/pa)^2 + \pi^2 a^2\mu_0\epsilon_0\epsilon_r m^2 y^2[m^2y^2F^2(y) - 2myF(y)G(y) + x^2G^2(y)]\tan\delta}{2\beta_2\{4x^2 - \pi^2 my(x^2 - m^2y^2)F(y)[myF(y) - 2G(y)]\}}$$

Since $$x^2 = m^2y^2 + (\epsilon_r - 1)k_0^2b^2 \quad . \quad . \quad . \quad . \quad (48)$$

he expression for α_2 can be written in terms of ϵ_r, m, and y. convenient form of the expression is

$$\alpha_2 = \frac{\alpha_1}{S} + \frac{\epsilon_r k_0^2 T}{2\beta_2 S}\tan\delta \quad . \quad . \quad . \quad . \quad (49)$$

here

$$S = \frac{a^2p^2\beta_2}{4\beta_1x^2y^2}\{4x^2 - \pi^2my(x^2 - m^2y^2)\mathrm{F}(y)[my\mathrm{F}(y) - 2\mathrm{G}(y)]\} \quad . \quad . \quad . \quad . \quad (50)$$

nd

$$T = \frac{\pi^2m^2a^2p^2}{4\beta_1x^2}[m^2y^2\mathrm{F}^2(y) - 2my\mathrm{F}(y)\mathrm{G}(y) + x^2\mathrm{G}^2(y)] \quad . \quad (51)$$

) RELATIONSHIP BETWEEN THE CAVITY Q-FACTORS AND THE ATTENUATION COEFFICIENTS

The stored energies and power losses required in eqn. (14) ill be evaluated from the fields existing in the cavity at resonance. he losses are calculated by the usual method of perturbing the olution for a loss-free cavity, and since the effects of the coupling oles, the end walls and the evanescent modes at the discon- nuity have already been eliminated, they need not be considered urther. It is therefore assumed that the end walls are perfect onductors and that evanescent modes need not be considered. his second assumption is equivalent to taking the lengths l', $'$ in the equivalent circuit of Fig. 2 equal to the physical lengths d in Fig. 1. In all the cases which have been examined, this fully justified. A check is available in that the distances l_1, can be derived from the experimental curves corresponding Fig. 3.

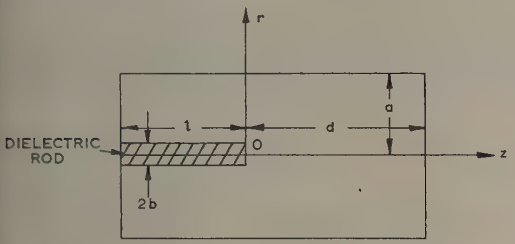

Fig. 9.—Co-ordinate system used in Section 8.

The fields in the cavity can therefore be represented sufficiently ccurately by H_{01} modes in the two sections. The co-ordinate ystem used is shown in Fig. 9. For $0 \leqslant z \leqslant d$

$$E_\theta = AJ_1(pr)\sin[\beta_1(d - z)] \quad . \quad . \quad . \quad . \quad (52)$$

$$H_r = jAY_1J_1(pr)\cos[\beta_1(d - z)] \quad . \quad . \quad . \quad (53)$$

here $Y_1 = \beta_1/\omega\mu_0$ is the wave admittance of the H_{01} mode in rcular guide.

(8.1) Total Stored Energy

The total energy stored in the cavity at resonance will be btained by an application of Slater's theorem, which relates hanges in resonant frequency to changes in stored energy. he frequency shift, $\delta\omega$, associated with a small reduction δd the length of the empty section can be calculated from eqn. (5). ater's theorem applied to this situation gives:

$$\frac{\delta\omega}{\omega} = \frac{\text{Difference between the magnetic and electric energies stored in the section}}{\text{Total stored energy in the cavity at resonance}}$$

$$= \frac{\delta W_m - \delta W_e}{W_1 + W_2} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (55)$$

where W_1, W_2 are defined in Section 3.2.

The changes δW_m and δW_e are readily calculated from the field components given by eqns. (52)–(53) evaluated for $z = d$ and are

$$\delta W_m = \frac{1}{4}\mu_0 Y_1^2|A|^2x2\pi\int_0^a rJ_1^2(pr)dr \,.\, \delta d$$

$$= \frac{\pi}{4}\mu_0 Y_1^2|A|^2a^2J_0^2(pa)\delta d \quad . \quad . \quad . \quad . \quad (56)$$

$$\delta W_e = 0 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (57)$$

Hence, $$W_1 + W_2 = \frac{1}{4}\pi\omega\mu_0Y_1^2|A|^2a^2J_0^2(pa)\frac{\delta d}{\delta\omega} \quad . \quad . \quad (58)$$

Now if l_1 and d_1 are assumed to be zero, eqn. (5) becomes

$$K^2\tan\beta_1 d = -\tan\beta_2 l \quad . \quad . \quad . \quad . \quad (59)$$

Hence, since d is reduced by δd,

$$-n^2\beta_1\delta d\sec^2\beta_1 d = -\frac{d}{d\omega}(K^2\tan\beta_1 d + \tan\beta_2 l)\delta\omega$$

$$= -\left(\frac{dK^2}{d\omega}\tan\beta_1 d + K^2 d\sec^2\beta_1 d\frac{d\beta_1}{d\omega} + l\sec^2\beta_2 l\frac{d\beta_2}{d\omega}\right)\delta\omega \quad (60)$$

Further, it has been shown[18] that, to a very good approximation,

$$K^2 = \beta_2/\beta_1 \quad . \quad . \quad . \quad . \quad . \quad . \quad (61)$$

i.e. $$\frac{dK^2}{d\omega} = \frac{1}{\beta_1}\frac{d\beta_2}{d\omega} - \frac{\beta_2}{\beta_1^2}\frac{d\beta_1}{d\omega} \quad . \quad . \quad . \quad . \quad (62)$$

Hence

$$\frac{\delta d}{\delta\omega} = \frac{1}{\beta_1}\left[\frac{1}{2}\left(\frac{1}{\beta_2}\frac{d\beta_2}{d\omega} - \frac{1}{\beta_1}\frac{d\beta_1}{d\omega}\right)\sin(2\beta_1 d) + d\frac{d\beta_1}{d\omega} + l\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2\beta_1 d\right)\frac{d\beta_2}{d\omega}\right] \quad (63)$$

and so

$$W_1 + W_2 = \frac{\pi\beta_1|A|^2a^2J_0^2(pa)}{4\omega\mu_0}\left[d\frac{d\beta_1}{d\omega} + l\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2\beta_1 d\right)\frac{d\beta_2}{d\omega} + \frac{1}{2}\left(\frac{1}{\beta_2}\frac{d\beta_2}{d\omega} - \frac{1}{\beta_1}\frac{d\beta_2}{d\omega}\right)\sin 2\beta_1 d\right] \quad . \quad . \quad . \quad . \quad (64)$$

The validity of the approximations used in the above derivation has been confirmed by evaluating $(W_1 + W_2)$ by the very tedious method of integrating over the volume of the cavity.

The expression for $W_1' + W_2'$ is obtained by substituting $d - \frac{1}{2}\lambda_1$ and $l + \frac{1}{2}\lambda_2$ for d and l, respectively.

The derivatives $d\beta_1/d\omega$, $d\beta_2/d\omega$ which appear in eqn. (64) will be evaluated in the next Section.

(8.2) Derivatives $\frac{d\beta_1}{d\omega}$, $\frac{d\beta_2}{d\omega}$

Since $$\beta_1^2 = \omega^2\mu_0\epsilon_0 - p^2$$

$$\beta_1\frac{d\beta_1}{d\omega} = \omega\mu_0\epsilon_0$$

i.e. $$\frac{d\beta_1}{d\omega} = \frac{k_0^2}{\beta_1\omega} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (65)$$

The expression for $d\beta_2/d\omega$ is not obtained in this simple way, since β_2 depends in a complicated manner on ϵ_r, a, b and k_0. From eqns. (8) and (9),

$$x^2 = (\epsilon_r k_0^2 - \beta_2^2)b^2$$

$$y^2 = (k_0^2 - \beta_2^2)a^2$$

where x and y are related by eqn. (10).

Hence

$$x\frac{dx}{d\omega} = \left(\frac{\epsilon_r k_0^2}{\omega} - \beta_2\frac{d\beta_2}{d\omega}\right)b^2 \quad . \quad . \quad . \quad . \quad (66)$$

$$y\frac{dy}{d\omega} = \left(\frac{k_0^2}{\omega} - \beta_2\frac{d\beta_2}{d\omega}\right)a^2 \quad . \quad . \quad . \quad . \quad (67)$$

and so

$$\frac{dy}{dx} = \frac{dy}{d\omega}\cdot\frac{d\omega}{dx} = \frac{\left(\frac{k_0^2}{\omega} - \beta_2\frac{d\beta_2}{d\omega}\right)x}{\left(\frac{\epsilon_r k_0^2}{\omega} - \beta_2\frac{d\beta_2}{d\omega}\right)m^2 y} \quad . \quad . \quad (68)$$

$$\beta_2\frac{d\beta_2}{d\omega} = \frac{k_0^2}{\omega}\left(y\epsilon_r\frac{dy}{dx} - \frac{x}{m^2}\right)\Big/\left(y\frac{dy}{dx} - \frac{x}{m^2}\right)$$

$$= \frac{k_0^2}{\omega}\left[1 + \frac{y(\epsilon_r - 1)(dy/dx)}{y(dy/dx) - (x/m^2)}\right] \quad (69)$$

Now

$$\frac{J_1(x)}{xJ_0(x)} = \frac{G(y)}{myF(y)}$$

where $F(y)$, $G(y)$ are defined in Section 7.2.

By straightforward differentiation and substitution for the Bessel derivatives it is found that

$$\frac{dy}{dx} = \frac{\pi^2 y[m^2y^2F^2(y) - 2myF(y)G(y) + x^2G^2(y)]}{x\left\{\pi^2m^2y^2[F^2(y) + G^2(y)] - 2\pi^2 myF(y)G(y) - 4\right\}} \quad (70)$$

and after some further algebraic manipulation:

$$\frac{y(dy/dx)}{y(dy/dx) - (x/m^2)} = \frac{T}{S} \quad . \quad . \quad . \quad . \quad (71)$$

where S, T are the functions defined in eqns. (50) and (51). Substitution in eqn. (69) gives

$$\beta_2\frac{d\beta_2}{d\omega} = \frac{k_0^2}{\omega}\left[1 + \frac{(\epsilon_r - 1)T}{S}\right] \quad . \quad . \quad . \quad (72)$$

(8.3) Losses in the Cavity

The method proposed in Section 3.2 makes it unnecessary to evaluate the total losses in the cavity, and it suffices to determine the losses in half-wavelength sections of the empty and dielectric-filled regions. This can be done by considering the Q-factors of such sections. Consider a half-wavelength section of the empty cavity, for which the Q-factor can be calculated from either of the expressions

$$Q_1 = \frac{\omega}{2d_1}\frac{d\beta_1}{d\omega} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (73)$$

or $$Q_1 = \frac{\omega \times \text{Total energy stored in a half-wavelength}}{\text{Power loss in a half-wavelength}}$$

$$= \frac{\omega W_1(\lambda_1/2)}{P_1 - P_1'} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (74)$$

where P_1, P_1' have been defined in Section 3.2 and $W_1(\lambda_1/2)$ is the energy stored in a length $\lambda_1/2$ of the empty cavity. This may be obtained as a special case of eqn. (64) by letting d eq
$\lambda_1/2$ and l equal zero,

i.e. $$W_1(\lambda_1/2) = \frac{\pi\beta_1|A|^2a^2J_0^2(pa)}{4\omega\mu_0}\frac{\lambda_1}{2}\frac{d\beta_1}{d\omega} \quad . \quad . \quad ($$

If the two expressions for Q_1 are equated, there results

$$P_1 - P_1' = \frac{\omega W_1(\lambda_1/2)}{Q_1} = \frac{\pi^2\alpha_1|A|^2a^2J_0^2(pa)}{2\omega\mu_0} \quad . \quad ($$

By a similar argument it may be shown that

$$P_2' - P_2 = \frac{2\alpha_2 W_2(\lambda_2/2)}{(d\beta_2/d\omega)} \quad . \quad . \quad . \quad ($$

where $W_2(\lambda_2/2)$ is the energy stored in a half-wave section of
filled cavity section. This must be evaluated for the situat
when the fields in the empty cavity section are given
eqns. (52)–(53), and is obtained by taking the difference betw
$W_1 + W_2$ evaluated for d, $l + \frac{1}{2}\lambda_2$ and $W_1 + W_2$ evalua
for d, l.

Hence

$$W_2(\lambda_2/2) = \frac{\pi\beta_1|A|^2J_0^2(pa)a^2}{4\omega\mu_0}$$

$$\times\left[\tfrac{1}{2}\lambda_2\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2 p_1 d\right)\frac{d\beta_2}{d\omega}\right] \quad ($$

and $$P_2' - P_2 = \frac{\pi d_2\beta_1 a^2|A|^2J_0^2(pa)\lambda_2}{4\omega\mu_0}$$

$$\times\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2\beta_1 d\right) \quad ($$

(8.4) Final Form of Eqn. (14)

The various quantities appearing in eqn. (14) have b
evaluated in Sections 8.1 and 8.3, and substitution of the res
from eqns. (64), (76) and (79) gives

$$\frac{\pi^2\alpha_2 a^2|A|^2J_0^2(pa)}{2\omega\mu_0}\left(\sin^2\beta_1 d + \frac{\beta_1^2}{\beta_2^2}\cos^2\beta\, d\right) - \frac{\pi^2\alpha_1 a^2|A|^2J_0^2(pa}{2\omega\mu_0}$$

$$= \frac{\pi\omega\beta_1 a^2|A|^2J_0^2(pa)}{4\omega\mu_0}\left\{\frac{1}{Q'}\left[(d - \tfrac{1}{2}\lambda_1)\frac{d\beta_1}{d\omega} + (l + \tfrac{1}{2}\lambda_2)\right.\right.$$

$$\times\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2\beta_1 d\right)\frac{d}{d}$$

$$+ \tfrac{1}{2}\sin(2\beta_1 d)\left(\frac{1}{\beta_2}\frac{d\beta_2}{d\omega} - \frac{1}{\beta_1}\frac{d\beta_1}{d\omega}\right)\left(\frac{1}{Q'} - \frac{1}{Q}\right)$$

$$\left.- \frac{1}{Q}\left[d\frac{d\beta_1}{d\omega} + l\left(\frac{\beta_1}{\beta_2}\cos^2\beta_1 d + \frac{\beta_2}{\beta_1}\sin^2\beta_1 d\right)\right]\right\} \quad . \quad ($$

This equation may be simplified by substituting for
$(d\beta_1/d\omega)$ and $(d\beta_2/d\omega)$ from eqns. (49), (65) and (72), resp
tively, and by defining

$$R = \sin^2(\beta_1 d) + \left(\frac{\beta_1}{\beta_2}\right)^2\cos^2(\beta_1 d) \quad . \quad . \quad ($$

The final result is

$$\epsilon_r RT\tan\delta = \frac{2\beta_2(S - R)d_1}{k_0^2} + \frac{1}{Q'}\left[\left(R - \frac{\beta_2}{\beta_1}\right)S + (\epsilon_r - 1)RT\right]$$

$$+ \frac{1}{\pi}\left(\frac{1}{Q'} - \frac{1}{Q}\right)\left\{\beta_2(d + Rl)S + (\epsilon_r - 1)\beta_2 lRT\right.$$

$$\left.+ \frac{1}{2}\left[\left(\frac{\beta_1}{\beta_2} - \frac{\beta_2}{\beta_1}\right)S + \frac{\beta_1}{\beta_2}(\epsilon_r - 1)T\right]\sin(2\beta_1 d)\right\} \quad ($$

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PROPAGATION MEASUREMENTS AT 3480 AND 9640 Mc/s BEYOND THE RADIO HORIZON

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SUMMARY

The paper gives an account of scatter-propagation measurements made at S- and X-band during the period May, 1957 to May, 1959. The work at S-band continued throughout the whole of this time, while the X-band measurements were made during the period of a year from June, 1958 to May, 1959.

The S-band transmitter, using a 3·480 Gc/s c.w. magnetron with a power output of 500 watts, was established at Start Point in Devon. Receiving terminals were set up at Wembley, Middlesex, and Witnesham, Suffolk, at distances of 173 and 247 statute miles, respectively, from the transmitter, the former being maintained in operation throughout the whole of the experimental period and the latter for a period of nine months from September, 1957 to June, 1958. Diurnal and seasonal variations in the median level of the received signal are discussed and a comparison is made of measurements taken simultaneously at Wembley and Witnesham. The distribution of the S-band fading rate as a function of level is studied and a limited amount of work concerned with the distribution of the duration of fades below a given level and the power spectrum of the detected signal is described.

The transmission path for the X-band system, which used a pulsed magnetron at a frequency of 9·640 Gc/s, also lies between Start Point and Wembley; the pulse length was 2 microsec and the pulse-repetition frequency 500 c/s, the peak power in the pulses being 180 kW. The X- and S-band links were operated together whenever possible, and a comparison is made of the median level and fading rate of the signals received simultaneously at the two frequencies over the same propagation path. In addition, a series of measurements to investigate the aerial coupling loss of the X-band system are described.

LIST OF SYMBOLS

f = Frequency.
f_0 = Frequency of transmitter.
k = R.M.S. amplitude of a Rayleigh distribution.
k_0 = R.M.S. amplitude of a Rayleigh distribution measured relative to a constant amplitude.
L = Transmission loss, dB.
L_c = Aerial-to-medium coupling loss, dB.
$N(r)$ = Number of crossings per second of a level r in a given sense.
r = Amplitude of a field quantity.
$\bar{t}(r)$ = Average duration of fades below a level r.
$\bar{t}_0$ = Average duration of fades below the median level.
ξ = Amplitude of a field quantity measured relative to the median value.
$\phi(r) = \int_0^r N(r)\,dr$.
$\psi(\tau)$ = Auto-correlation function.
σ = Standard deviation of a normal distribution.
θ = Scattering angle.
$\omega(f)$ = Power spectrum.

(1) INTRODUCTION

In a previous paper[1] an account was given of an experimental study of beyond-the-horizon propagation using a 500-watt c.w. transmitter at a frequency of 3·480 Gc/s over a 173-mile path between Start Point, Devon, and Wembley, Middlesex, for the period May, 1956 to April, 1957. This study has been continued and expanded, and in this paper a summary is given of the propagation data collected up to May, 1959. In addition to the receiving equipment at Wembley which has been in operation throughout the programme, a second receiving station was established at Witnesham, Suffolk, for the period September, 1957 to June, 1958. Witnesham lies on the extension of the line through Start Point and Wembley and is 247 miles from the transmitter at Start Point. Simultaneous signal-level recordings at the two sites enabled an estimate to be made of the average attenuation rate of the scatter-type signal at S-band.

The transmissions from Start Point have also been monitored at Ditton Park by the Radio Research Station of the D.S.I.R. as part of a programme to study tropospheric propagation. Ditton Park is situated near Slough and is only slightly off the Wembley–Start Point line at a distance of 13 miles from Wembley. Apart from aerial size, the receiving equipment at Slough was identical with that used at Wembley and Witnesham. Information was exchanged with the D.S.I.R., and, in particular, a comparison extending over several months was made of the variations in hourly median signal levels at Slough and Wembley.

In addition to the work at S-band, a second propagation link, operating at 9·640 Gc/s, was established over the same path between Start Point and Wembley in June, 1958. This used a pulsed radar transmitter with a pulse length of 2 microsec, a pulse-repetition frequency of 500 c/s and a peak power output of 180 kW. Whenever possible the S- and X-band links have been operated simultaneously, and comparisons of hourly median transmission loss extending over the period of a year, together with simultaneous measurements of fading rate at the two frequencies, are given in the paper.

Table 1 summarizes some of the more important parameters of the links.

(2) DESCRIPTION OF EQUIPMENT

(2.1) S-Band

Three separate receiving equipments have been used during the programme; two of these were installed on a 200 ft tower at Wembley and the third at Witnesham. The second Wembley receiver was installed originally to enable the experiments on double diversity reported in Reference 1 to be made and was subsequently retained as a standby receiving channel. Details of the siting of the equipment at Start Point and Wembley, together with a brief outline of the characteristics of the transmitter and Wembley receivers are given in Reference 1. The receiver characteristics given there apply equally well to the Witnesham equipment.

(2.2) X-Band

The details of the X-band equipment necessary for the interpretation of the propagation data are given below.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.
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Table 1

	S-band		X-band
	Wembley	Witnesham	Wembley
Angular distance, m.r. ..	26	41	26
Yearly median system transmission loss, dB	151	162	153
Aerial coupling loss, dB (estimated)	3	5	8
Transmission loss between isotropic aerials, dB	221	230	232
Calculated transmission loss between isotropic aerials (Reference 2), dB	222	236	234
Standard deviation of the distribution of hourly medians, dB	5	5	7
Average seasonal variation in monthly median values of system transmission loss, dB	8	10	17
Average yearly range in the hourly median transmission loss, dB	60	51	54
Average diurnal range in hourly median transmission loss, dB	14	13	15
Average monthly variation in hourly median transmission loss, dB	28	24	26
Median value of fading rate at median level, c/s	2·4	—	4·8

(2.2.1) Transmitter (Start Point).

The transmitter is housed in a trailer with the paraboloid mounted on the roof, and is adjacent to the trailer containing the S-band equipment. The characteristics of the equipment are as follows:

Frequency	9·640 Gc/s
Radiation	2 microsec pulses
Pulse-repetition rate	500 c/s
Valve	CV2412
Peak power output	180 kW
Aerial	8 ft-diameter 120° paraboloid
Height of centre of aerial above ground	16 ft
Aerial feed	Pyramidal horn
Aerial gain	45 dB
E-plane half-power beam width	1·0°
H-plane half-power beam width	1·0°
Polarization	Horizontal

(2.2.2) Receiver (Wembley)

The X-band equipment, like that for S-band, is housed at the top of a 200 ft tower at Wembley and shares a paraboloid with the second S-band receiver. The characteristics of the receiver are as follows:

Aerial	8 ft-diameter 120° paraboloid
Height of aerial above ground	190 ft
Aerial feed	Dual X- and S-band feed
Aerial gain	42·1 dB
E-plane half-power beam width	1·2° (estimated)
H-plane half-power beam width	1·2° (estimated)
Polarization	Horizontal
Bandwidth	2 Mc/s
Noise factor	11·7 dB

(2.3) Recording Equipment

Three methods of recording signal level have been used:

(*a*) A low-speed pen recorder having a paper speed whi could be alternatively 12 or 3 in/hour and a time-constant about 1 sec.

(*b*) A high-speed two-pen recorder with a level response to 60 c/s and paper speeds of 1, 5, 25 and 100 mm/s.

(*c*) A signal sampling machine.

At Wembley all three methods were available for the S-ba signal but only (*a*) and (*b*) for X-band; at Witnesham only low-speed pen recorder was used. The signal sampling machi which enables the amplitude distribution of the detected S-ba signal over any desired period to be measured, is briefly describ in Reference 1.

At X-band it was necessary to feed the output of the vid amplifier into a pulse-stretching circuit, which converted t pulses into a varying direct current whose amplitude follow those of the successively arriving pulses and had a mean val large enough to operate a conventional pen recorder. Since t pulse-repetition frequency was large compared with the fadi rate of the X-band signal, the waveform from the pulse stretch mirrored fairly closely the variations in signal amplitude whi would have been obtained under c.w. conditions.

(2.4) Calibration of the Receivers

(2.4.1) S-Band.

The original receiver at Wembley was calibrated at t beginning of the programme in 1956 using a small batter operated transmitter of known output over a 16·75-mile line-o sight path from Windsor. In 1957, when all three receivers we in operation, another calibration, this time covering all thr receivers, was carried out using a similar method but over pa lengths of about 1·25 miles. The total spread in the measur sensitivity of the three nominally identical receivers was 1·5 dB a variation which could mainly be accounted for by the differen between the mixer crystals. The comparatively small spre between the measurements on the three receivers gives sor confidence in the accuracy of these results, and the median-lev measurements based on the original Windsor calibration ha therefore been modified to conform with the later calibratic The correction was, however, only about 1 dB.

(2.4.2) X-Band.

The X-band receiver was calibrated by feeding the output o pulsed r.f. signal generator giving 2 microsec pulses at a pul repetition frequency of 500 c/s directly into the aerial feed.

(3) TRANSMISSION PATHS

The locations of the various terminals are shown in Fig. Brief descriptions of the sites at Start Point and Wembl together with profiles of the path in the immediate neighbo hood of these terminals, are given in Reference 1. The site Witnesham is on farm land with the path of the signal in t immediate neighbourhood of the aerial lying across open fiel A profile of the transmission path near this terminal is giv in Fig. 2. The radio horizons under standard radio conditic at Start Point, Wembley and Witnesham are, respectively, 17 and 9·5 miles, and the angular distances[2] of the Start Poir Wembley and Start Point–Witnesham paths are 25·9 a 41·2 millirad.

(4) S-BAND MEDIAN-LEVEL MEASUREMENTS

The signals received over a tropospheric scatter link characterized by rapid fading over a large range, together wit

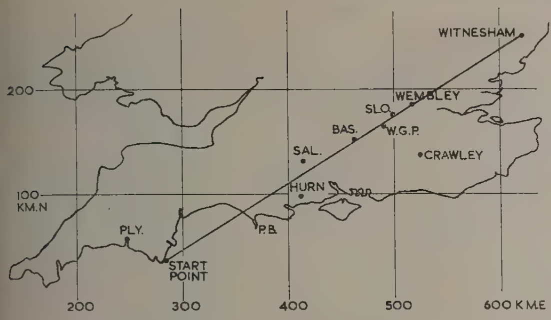


Fig. 1.—Map of the propagation path.

Start Point: 50°13′36″N, 3°39′38″W, NG 815375.
Wembley: 51°33′48″N, 0°17′45″W, NG 181864.
Witnesham: 52° 7′36″N, 1°11′ 4″E, NG 180524.

BAS.	Basingstoke.	SAL.	Salisbury.
P.B.	Portland Bill.	SLO.	Slough.
PLY.	Plymouth.	W.G.P.	Windsor Great Park.

Scale: 1 in = 40 miles (approximately).

8 hours before returning to its original value, although one exceptional period occurred in January, 1959, in which a high median level, some 25 dB above the average for the rest of the month, persisted for 24 hours. Over three years a sufficient number of periods of ducting have been observed to suggest a seasonal dependence of the atmospheric conditions necessary to produce such signals. They seem to occur far more frequently in summer and autumn, and only very rarely in winter; in fact, the exceptional period in January, 1959 mentioned above is the only time such a signal has been observed during the winter months.

In Reference 1 it was reported that the times of occurrence of the maximum and minimum values of the hourly median transmission loss within a period of 24 hours for the period May, 1956 to April, 1957 were distributed fairly uniformly throughout the day, suggesting that there was no systematic diurnal variations in median level. An increase in the rate of accumulation of data during the last two years of operation made it possible to study this question in more detail. For this purpose the high-level ducted-type signals, defined here as ones

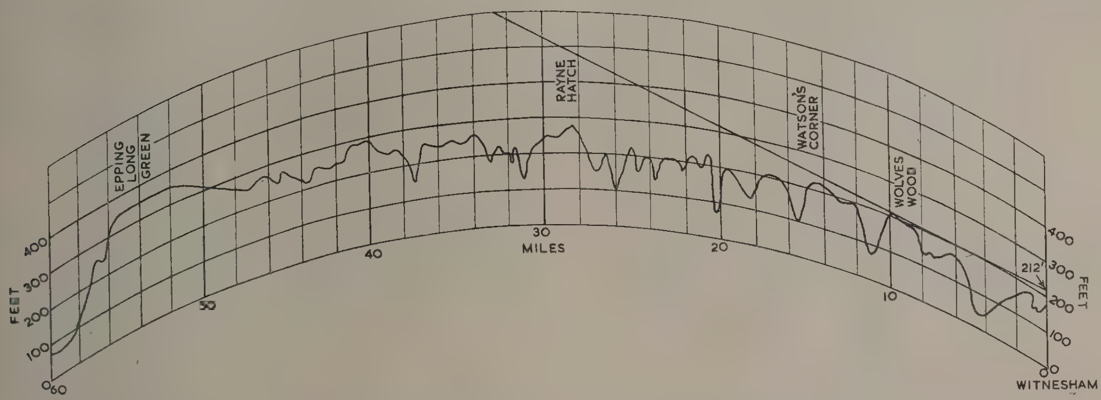


Fig. 2.—Part of profile between Start Point and Witnesham.

ower variation in the average level. The measurement of such gnals therefore resolves itself into two parts: first, a determinaon of the long-term variations in the average level, and secondly, e determination of the distribution of the amplitude about the ean over a relatively short period. The former will be dealt ith in this Section.

(4.1) At Wembley

The long-term variations in the average level of the received gnal have been measured, for the most part, in terms of the edian value of the system transmission loss, L, over the period f one hour, where L is defined as the ratio, expressed in decibels, f the power transmitted to that received. The total recording me at Wembley from May, 1956 to May, 1959 was about 500 hours, and during this period the measured values of L aried from 108·5 to 171·0 dB, a range of 62·5 dB. The conusions drawn in Reference 1 about the variability of L on the asis of the results of one year apply substantially unaltered to e whole period of operation, the mean values of the diurnal nd monthly variations in L for the 3-year period being, respecvely, 13·0 and 29·2 dB. In the last two years there has been erhaps a slight increase in the frequency of occurrence of the ucted' type of signal in which the median level increased quite uddenly by up to 30 dB. Such a signal usually lasted from 1 to

for which $L < 135$ dB, were separated out from the rest of the data, and separate analyses were carried out on the two sets of results. For the more usual data ($L > 135$ dB) plots were made of the monthly averages of the median values of the transmission loss for each hour of the day; at least 7 values were available to average and in some months as many as 15. The curves showed no well-marked common features, and, in general, the variation throughout the day was quite small. On the other hand, a histogram of the times of occurrence of high signal ($L < 135$ dB) showed clearly that the ducted type of signal occurred far more frequently between 2200–0600 hours with a peak period just after midnight. The high signals were recorded for approximately 2% of the total time of observation. It appears, therefore, that the basic mechanism responsible for the transmission of S-band signals beyond the horizon is, on the whole, equally effective throughout the day. On the other hand, the special atmospheric conditions which lead to greatly enhanced signals occurred mainly for the few hours on either side of midnight.

The distribution of the hourly medians over a month or longer period was found, in general, to be approximately Gaussian. Cumulative distributions of hourly median values, plotted on arithmetic-probability paper, for the best and worst months and for the whole year are shown in Fig. 3 for 1958. The monthly

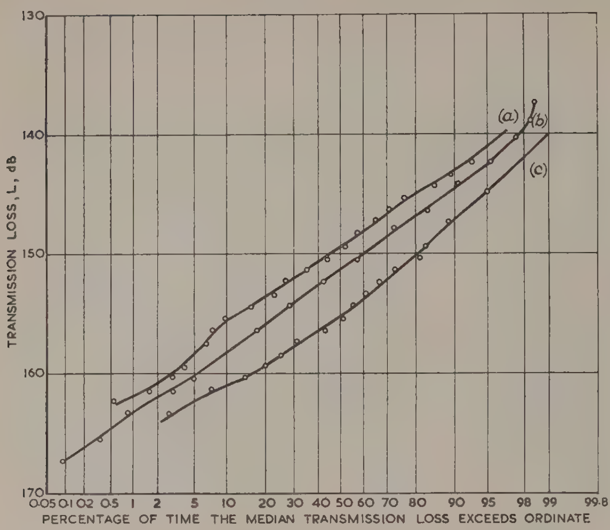


Fig. 3.—Distribution of the hourly median S-band transmission loss for the Start Point–Wembley path for 1958.

(a) Best month (August).
(b) Year.
(c) Worst month (January).

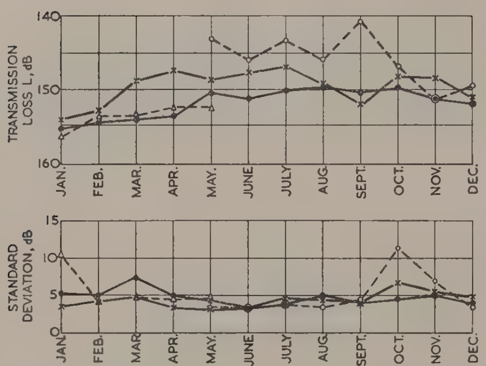


Fig. 4.—Monthly median and standard deviation of the S-band transmission loss for the Start Point–Wembley path.

○— —○ 1956.
×——× 1957.
●——● 1958.
△— —△ 1959.

median values and standard deviations over the period May, 1956 to May, 1959 are plotted in Fig. 4. A seasonal trend in the median values is evident, the signal being, on the whole, higher from May to October than in the rest of the year. The loss is greatest in January and February, and the difference in median level between these and the summer months is, on the average, about 7–8 dB. There is, however, a range of about 5 dB between the average summer levels for 1956, 1957 and 1958. The monthly spread in hourly medians as measured by the standard deviation of the distribution usually varies from 4 to 6 dB with a mean value of about 4·5 dB. Months in which the standard deviations differed significantly from the average usually contained relatively long periods of ducted-type signals. The average yearly median value of the system transmission loss is about 151 dB.

(4.2) At Witnesham

The signal level at Witnesham was expected to be only a f[ew] decibels above the receiver threshold for quite a large proport[ion] of the time. Because of this and the fact that the receiver w[as] left unattended for long periods, an automatic search circ[uit] was included in addition to the normal automatic freque[ncy] control. As long as the signal was above the threshold [the] search system was prevented from working and the norm[al] automatic frequency control operated. However, if the sig[nal] was lost for more than 12 min, the search system, which co[n]trolled the local-oscillator frequency by varying the reflec[tor] voltage on the klystron, came into operation and continued u[ntil] the signal was again picked up. This worked extremely w[ell] and, in spite of the fact that the receiver was frequently work[ing] quite close to its noise level, comparatively few periods [of] recording were lost.

Since the only recording equipment at Witnesham was a lo[w-]speed pen recorder with a maximum paper speed of 12 in/ho[ur] the amount of quantitative information which could be obtai[ned] about the signal was more restricted than at Wembley. It w[as] limited to the estimation of the median level by eye from [the] low-speed chart. A suitable length of chart for quick a[nd] repeatable estimates was found to be 6 in, and the basic per[iod] of analysis was therefore taken to be ½ hour rather than 1 ho[ur] as at Wembley. During the period from September, 1957 [to] June, 1958 about 950 hours of simultaneous recordings [at] Wembley and Witnesham were made. The measurements w[ere] spread throughout this period as indicated in Table 2, a[nd]

Table 2

Month	Sept.	Oct.	Nov.	Dec.	Jan.	Feb.	March	April	May	Ju[ne]
Number of hours	43	80	98	—	22	79	123	203	174	13[...]

included several periods of three or more days continu[ous] recording.

Although the Witnesham programme is clearly considerab[ly] smaller than that at Wembley, sufficient information was ava[il]able to give a resonably accurate picture of the propagati[on] throughout the year and to enable a meaningful comparison [of] median-level trends at the two sites to be made.

The general picture of signal variability in median level [is] substantially the same as at Wembley and is summarized [as] follows:

(a) The total measured variation in level over the peri[od] from September, 1957 to June, 1958 was 51 dB.

(b) The hour-to-hour variation was usually of the order [of] 1–2 dB.

(c) A few periods of high signal level of the type describ[ed] for Wembley were recorded. These usually coincided wi[th] similar signals at Wembley, although the extent of the sudd[en] increase in level might differ by about 10 dB at the two sit[es].

(d) The average diurnal and monthly variations we[re] respectively, 13·1 and 24·2 dB.

(e) The average changes in level within 6-hour periods a[re] given in Table 3, the very high signals having been omitted.

Table 3

Period	0000–0600	0600–1200	1200–1800	1800–240[0]
Average change, dB	6·3	6·7	6·8	6·8

As at Wembley, the signal appeared to be equally variable throughout the day.

(*f*) If the periods of ducted-type signal, which occurred most frequently around midnight, are omitted there appears to be no systematic variation in level with the time of day.

(*g*) The distribution of the half-hourly median transmission loss over the period of a month or longer was approximately Gaussian with a standard deviation of about 4–5 dB. The distribution of all the measurements taken over the period September, 1957 to June, 1958 is shown in Fig. 5. The

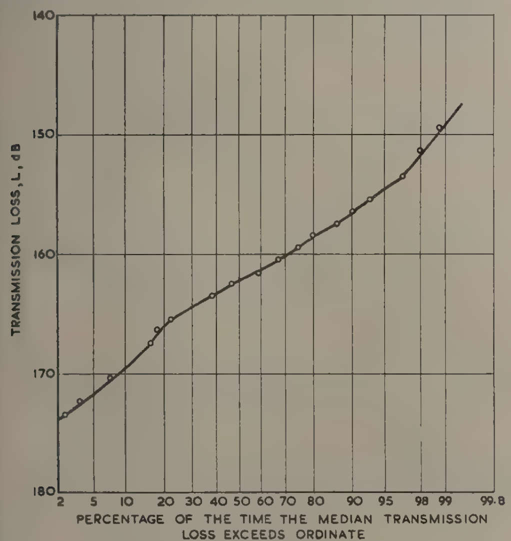


Fig. 5.—Distribution of the half-hourly S-band median transmission loss for the Start Point–Witnesham path.

median value of the half-hourly medians is 160·7 dB and the standard deviation is 4·9 dB.

(*h*) Although the amount of data in some months is too small to give a very accurate estimate of the monthly median level, there is evidence of the same seasonal trends as were found at Wembley, the difference in level between the best and worst month being about 10 dB.

(*i*) It is of interest to note that, although the measured yearly median value of the transmission loss for the Start Point–Wembley link was in good agreement with the value calculated from Reference 2, the corresponding calculated figure for the Start Point–Witnesham link was 6 dB greater than the measured value.

(4.3) Comparison of Wembley, Witnesham and Slough Results

The main object in establishing the second receiver site at Witnesham was to compare the median signal levels received simultaneously at two sites separated by a fairly large distance—in this case 74 miles—under as many conditions as possible, and to study the distribution of the difference in level between the two signals. From this an idea of the average attenuation rate of a tropospheric scatter signal at S-band could be obtained. To avoid systematic errors in the comparison arising from different methods of measuring signal level at the two sites, low-speed charts, rather than the signal sampling recorder, were used to estimate half-hourly median levels for Wembley.

Fig. 6 shows the variations in the half-hourly medians at Wembley and Witnesham for a period of 4 days' continuous recording during November, 1957. The relative behaviour of the signals at the two sites indicated by this Figure was the more

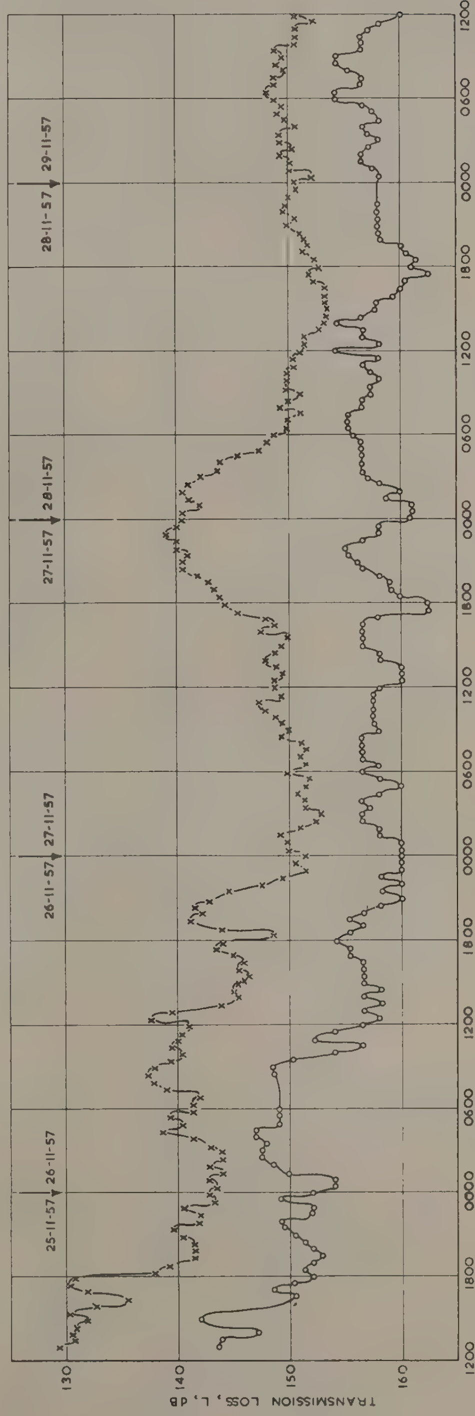


Fig. 6.—Comparison of the S-band median transmission loss for the Start Point–Wembley and Start Point–Witnesham paths for November, 1957.
×—×—×— Wembley.
○—○—○— Witnesham.

usual type observed. On the whole the trends in the two signals are similar, but there is little correlation between the detailed fluctuations. On occasions, however, the Witnesham signal was equal to, or even higher than, that at Wembley, while at other times it was as much as 30 dB lower. The periods of approximately equal signal strength resulted from either an enhancement at Witnesham or a decrease in the Wembley signal with the general level at the other site remaining substantially the same. This suggests that although, in general, the atmospheric conditions over Southern England are sufficiently uniform to make the behaviour of the effective scattering volumes for Wembley and Witnesham approximately the same, disturbances can occur which are sufficiently localized to affect one of the signals while leaving the other almost unaltered.

It is of interest to compare the diurnal variations in median level at the two sites. A summary of the results is given in Table 4.

It is seen that, with a few exceptions, the variations within a day are comparable. The average diurnal variations at Wembley and Witnesham are both about 13 dB.

A histogram of the difference between the half-hourly medians at Wembley and Witnesham is shown in Fig. 7; the differences range from the Witnesham signal being 7 dB above to 33 dB below that at Wembley. The distribution of the difference is approximately normal with a median value of 11·7 dB and a standard deviation of 5·8 dB. This corresponds to an average attenuation rate of 0·16 dB/mile.

The monthly median values and standard deviations for Wembley and Witnesham are plotted in Fig. 8, the Wembley results being derived from those periods in which there was simultaneous recording at Witnesham. This Figure again illustrates the fact that, for the most part, the general trends in the signal level at the two sites were similar.

Apart from the use of 3 ft 9 in instead of 8 ft-diameter paraboloids, the D.S.I.R. equipment at Slough was similar to that at Wembley, the signal level being recorded on a low-speed chart. Records covering about 250 hours and spread over a period of about six months were provided by the D.S.I.R., and the levels at Wembley and Slough were compared on the basis of half-hourly medians. Fig. 9 shows the results of this analysis for a period of four days in November, 1957. The graphs are typical of all periods examined. The average difference in absolute levels is very much what would be expected on the basis of aerial-gain differences, while the variations at the two sites follow one another quite closely. Fig. 9 should be compared with Fig. 6, which shows a comparison over the same

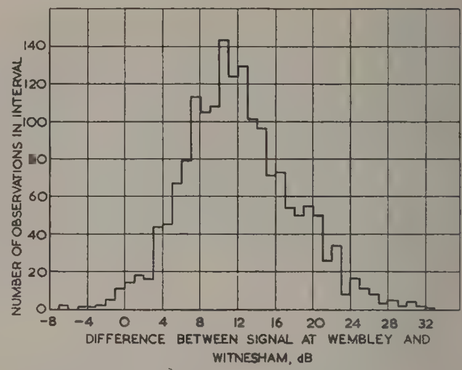


Fig. 7.—Histogram of the difference between the S-band med[ian] signal levels at Wembley and Witnesham.

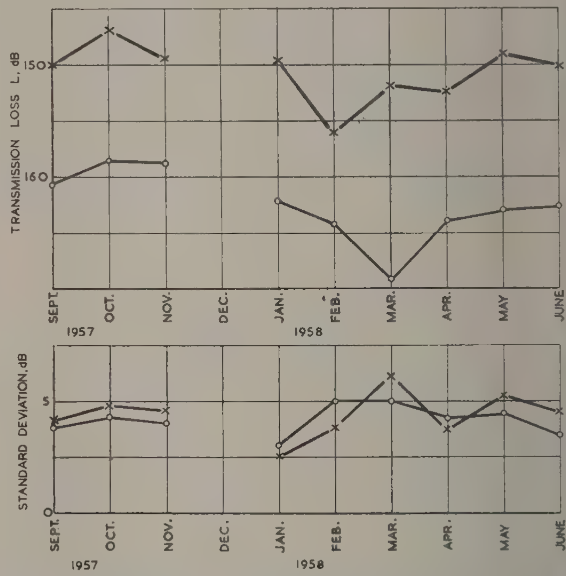


Fig. 8.—Comparison of the monthly S-band median transmission lo[ss] and standard deviation for the Start Point–Wembley and Sta[rt] Point–Witnesham paths.

×——× Wembley. ○——○ Witnesham.

Table 4

Test period number	Diurnal variations in median level 1		2		3		4		5		6		7		Mean diurnal variation	
	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.	Wit.	Wem.
	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB	dB
1957																
Sept. ..	14·9	13·0	10·6	15·4											12·8	14·2
Oct. ..	38·8	36·0	15·6	16·8											27·2	26·4
Nov. ..	14·5	14·7	6·1	15·2	9·3	11·0	8·3	5·3							9·6	11·6
Dec. ..	No measurements made														—	—
1958																
Jan. ...	7·5	7·9													7·5	7·9
Feb. ..	15·2	12·9	8·1	14·2	7·7	11·7									10·3	12·9
March ..	9·0	12·4	11·6	9·7	9·3	13·2	11·3	6·9	15·4	4·0	7·6	7·0			10·7	8·9
April ..	12·2	11·5	9·0	12·7	12·0	9·2	8·6	13·5	11·2	7·0	15·5	4·2	6·7	7·0	10·7	9·3
May ..	15·7	17·1	13·6	20·5	15·7	13·5	15·7	17·5	16·5	10·2	18·3	10·0	16·7	13·3	16·0	14·6
June ..	16·3	20·5	13·5	13·7	25·5	40·0	13·5	9·4	7·9	10·9	11·5	9·5			14·7	17·3

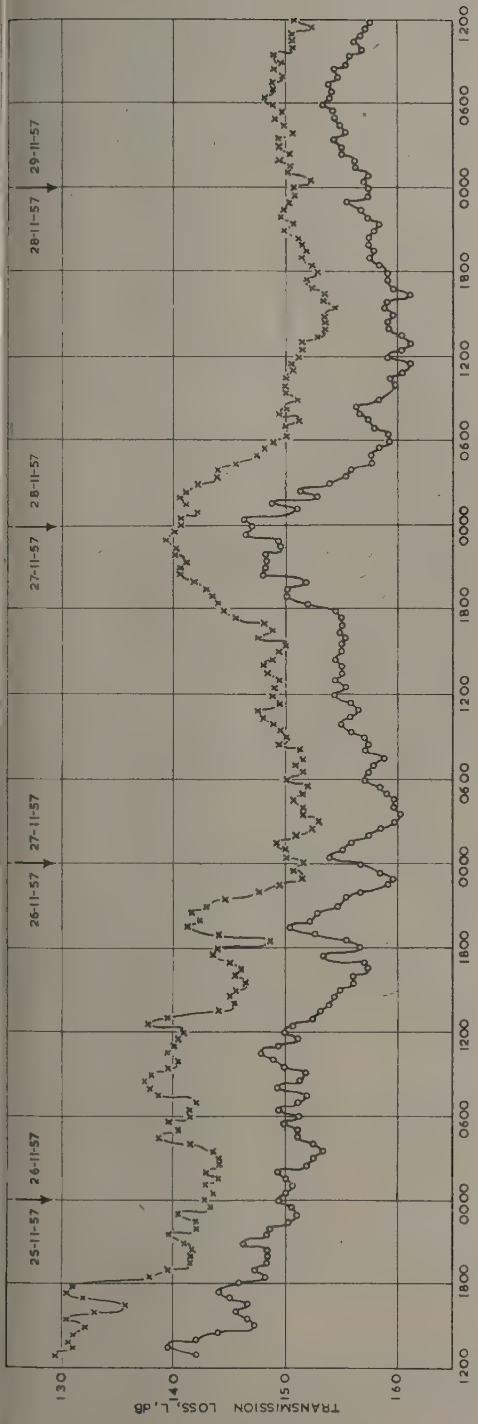


Fig. 9.—Comparison of the S-band median transmission loss for the Start Point–Wembley and Start Point–Slough paths for November, 1957.

×—×—× Wembley.
○—○—○ Slough.

period between Wembley and Witnesham. Thus, while at receiver sites separated by 74 miles it is usually only the general trends which are similar, at a distance of about 10 miles, provided that the angular path distances are comparable, there appears to be quite a strong correlation, at least for a large proportion of the time, between the detailed variations in median level.

(4.4) Aerial Siting Tests at S-Band

A short series of measurements were made at S-band to investigate the effect of aerial siting. A receiver, identical with that used at Wembley, was installed at Kenton, Middlesex, on a flat roof 30 ft above the gound and at a distance of 1·5 miles from the Wembley site. The Kenton terminal was shadowed by Harrow-on-the-Hill, which reduced the receiver horizon distance to 1·8 miles and increased the angular distance of the path from 25·9 millirad for the Start Point–Wembley link to 54 millirad. In addition, the scattering centre for the Start Point–Kenton path was about 40 miles nearer the receiver terminal and about 2 600 ft higher than for the path terminating at Wembley. The two links therefore made use of energy scattered from quite distinct regions of the atmosphere. The difference in median signal level to be expected is predicted, for example by the formula of Reference 2, to be about 15·7 dB.

About ten days' simultaneous recording, spread over a period of about six weeks, were made at Wembley and Kenton, and the levels were compared on the basis of half-hourly medians estimated from low-speed pen records of the received signals. Little correlation was observed between the variations in the median levels at the two sites, indicating that, in general, the behaviour of the atmosphere cannot be regarded as uniform over distances of the order of 40 miles. Four periods of ducting, each lasting for several hours, were observed during the tests. In two of these, ducted signals were observed at both sites, the times of occurrence and duration of the high signal being approximately the same at both terminals; on the other occasions ducting was present once at Wembley and once at Kenton, the average level at the other receiver remaining substantially unaltered. In the periods of simultaneous ducting at Wembley and Kenton the increase in level at the latter site was several decibels lower than at Wembley. The distribution of the measured differences in half-hourly medians at the two sites is approximately normal with a median value of 16 dB, agreeing quite well with the predicted value, and a standard deviation of about 4·3 dB.

(5) FAST FADING DURING PERIODS OF HIGH SIGNAL ($L < 135$ dB)

A detailed investigation of the amplitude distribution of the signal over the period of one hour under normal conditions ($L > 135$ dB) was presented in Reference 1, and was found in most cases to be similar to that of the amplitude of a constant vector plus a Rayleigh vector. Provided that the amplitude of the constant vector does not exceed the r.m.s. value of the Rayleigh component, a plot of such a distribution on Rayleigh graph paper is approximately a straight line of slope somewhat greater than unity, the value corresponding to the pure Rayleigh case. In this Section some measurements are reported of the signal-amplitude distribution under conditions of very high signal ($L < 135$ dB).

Relatively few complete amplitude distributions under high-signal-level conditions were obtained, because such signals occurred most frequently around midnight when the receiver was operating unattended. The gain of the receiver, set to accommodate the signal towards the end of the afternoon, was too high and resulted in loss of information from the signal

sampling machine. On the occasions when ducting was anticipated and the gain of the receiver was set to an appropriate low value, the amplitude distributions were examined over 15 min rather than the usual period of one hour.

Most of the 15 min amplitude distributions measured could be represented on Rayleigh paper, at least over the 10–90% range, reasonably well by straight lines. Several of the distributions, together with median values of the transmission loss, L, and the mean slopes, S, of the distribution curves are shown in Fig. 10. No absolute meaning is to be attached to the abscissa scale of this Figure. It is seen that the slopes tend, on the average, to be less than unity, the lowest value recorded being 0·6. This is in contrast to the distribution under normal conditions wl the slopes were unity or greater. The overall variations in le during the high-signal periods were therefore, on the wh greater than during normal propagation conditions.

It is interesting to note that Rayleigh-paper amplitude pl which, over a large part of the range, are straight lines of sl less than unity, can be obtained by combining n vectors (n i small integer) of constant amplitude but with phases which uncorrelated and uniformly distributed over 360°. A situat corresponding to this could arise for a link if the signal at receiver arose mainly from reflections at a small number elevated discontinuity layers. The usual scatter signal as

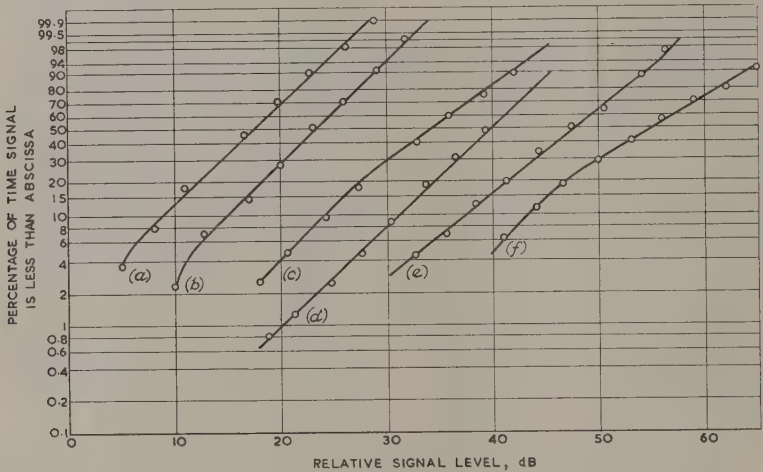


Fig. 10.—15 min S-band signal amplitude plots for a high-signal period at Wembley.

(a) $L = 129{\cdot}8$ $S = 0{\cdot}87$.
(b) $L = 129{\cdot}9$ $S = 0{\cdot}95$.
(c) $L = 125{\cdot}4$ $S = 0{\cdot}70$.
(d) $L = 133{\cdot}9$ $S = 0{\cdot}93$.
(e) $L = 127{\cdot}6$ $S = 0{\cdot}78$.
(f) $L = 124{\cdot}1$ $S = 0{\cdot}60$.

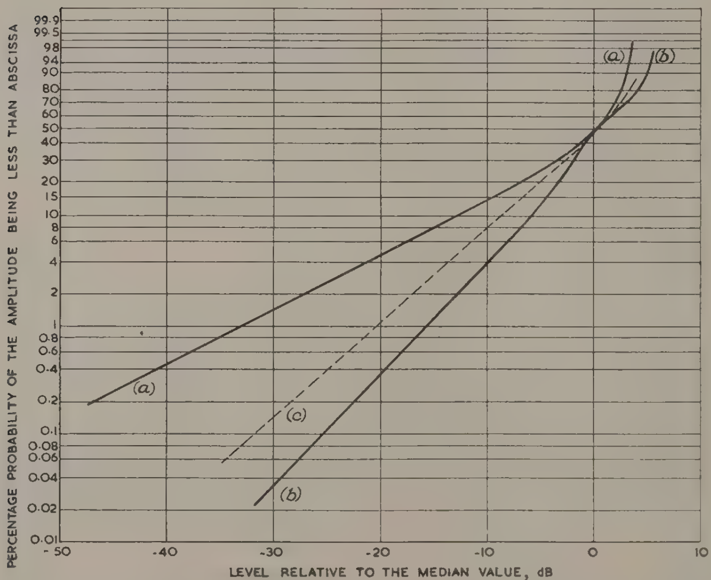
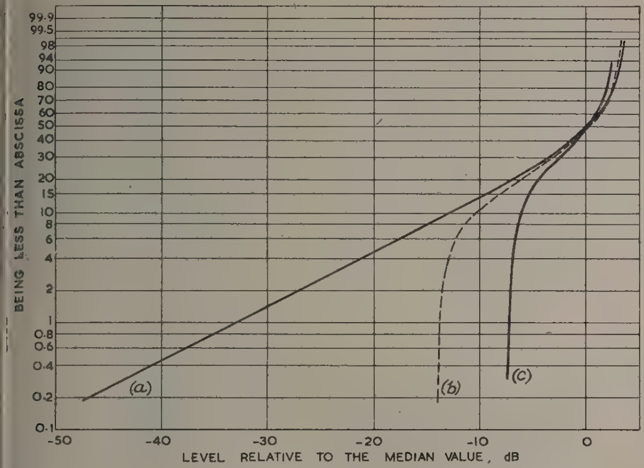


Fig. 11.—Amplitude distribution of the sum of a small number of randomly-phased vectors of equal magnitudes.

(a) $n = 2$ $S = 0{\cdot}52$.
(b) $n = 3$ $S = 1{\cdot}05$.
(c) $n = 4$ $S = 0{\cdot}90$.

n = Number of vectors.



ig. 12.—Amplitude distribution of the sum of two randomly phased vectors of different magnitudes.

(a) $m = 1$.
(b) $m = \frac{3}{4}$.
(c) $m = \frac{1}{2}$.

m = Relative magnitude of the two vectors.

iated with atmospheric turbulence would presumably still be resent, but its median value would be much lower than the mplitude of the reflected components. The distributions for = 2, 3 and 4 are shown in Fig. 11. It is clear from the graphs hat the distribution tends fairly rapidly to the Rayleigh distribuon as n is increased. For $n = 2$ and 4 the curves are straight nes up to the 70% level with slopes of 0·52 and 0·9, respecvely. The $n = 3$ curve is a straight line up to the 30% level ut then shows a pronounced kink. The slopes of all three urves become infinite at a few decibels above the median values ince the resultant amplitude cannot exceed the sum of the mplitudes of the component vectors.

The experimental results suggest, on the whole, the 2-vector ase, although the measured slopes, S, are never as small as the heoretical value. However, the slope for the case $n = 2$ can e altered by varying the relative amplitude, m, of the two vectors. 'heoretical plots for $m = \frac{1}{2}, \frac{3}{4}$ and 1 are shown in Fig. 12. It clear that quite a small difference between the amplitude of he two vectors could produce an appreciable change in the slope f the distribution. A signal consisting of two phase-indepenent components would appear to be a possible explanation of t least some of the results obtained under high signal conditions.

The only available information about the upper atmospheric onditions was contained in the Aerological Reports of the Air Ministry. Unfortunately none of the stations at which airorne measurements are made is particularly near the propagaion path. The two nearest are at Camborne and Crawley, where observations are made only twice a day and then not in ufficient detail to pick out relatively thin discontinuity layers. Iowever, from the available information a few M-profiles were lotted for periods of very high signal. In several of these there vas evidence of discontinuity layers in the height range 000–5 000 ft.

(6) FADING-RATE MEASUREMENTS AT S-BAND

The fading rate at a given signal level is defined as the average umber of times per second the envelope of the received signal rosses this level with a positive slope. Some measurements at he median, obtained from high-speed pen records, together with their correlation with the wind speed at a height of 1 500 m at the centre of the Start Point–Wembley path, were discussed in Reference 1. Some further work on this problem has been carried out with fading-rate estimates made both by a direct count of the number of crossings of a given level from a high-speed chart and automatically by using an electronic counter. The former method was tedious but had the advantage that any peculiarities in the record, such as amplitude modulation due to an aircraft in the beam, could easily be detected and discounted; in the latter method it was essential that the sampling period contained no aircraft effects if an over-estimate of the fading rate was to be avoided. A typical high-speed record is shown in Fig. 16 of Reference 1.

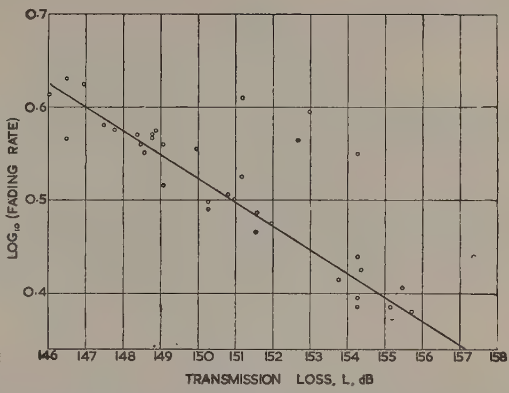


Fig. 13.—Variation of the logarithm of the S-band fading rate with transmission loss for the Start Point–Wembley path.

(6.1) Fading Rate at the Median Level

A considerable increase in the number of measurements at the median level has led to a revision of the estimates of average fading and spread quoted in Reference 1. The measured values varied from 0·06 c/s up to a maximum, observed under conditions of gale-force wind, of 14·3 c/s. The majority of the results were, however, in the range 0·5–6 c/s. The mean value of all the measurements was 2·49 c/s and the standard deviation was about 1·5 c/s. The correlation coefficient between the fading rate and the component of wind velocity normal to the propagation path for the larger sample was 0·49.

In addition to the above measurements, which were taken at a specific time of day to coincide with wind-speed measurements near the centre of the path, several days were spent in recording the fading rate at regular intervals throughout the period 0800–1600 hours. In all cases the rate varied appreciably throughout the day, a typical range being 1–4 c/s, and, in general, there appeared to be little correlation with the diurnal variations in median level. Two periods were, however, recorded in which the variations in fading rate and median level followed one another quite closely; in each case weather charts showed that a front was across the transmission path and advancing towards the receiver from the west. One of these results, taken on 1st July, 1958, is shown in Fig. 13 with the logarithm of the fading rate plotted against the transmission loss. Most of the points are seen to be approximately on a straight line. A similar result was found for the second period, although the slopes of the lines in the two cases differed considerably. It seems, therefore, that, although in general the fading rate and signal level are uncorrelated, conditions, perhaps associated with

the presence of fronts across the path, can arise in which the two quantities are closely correlated over periods of several hours.

(6.2) Fading Rate as a Function of Level

The number of times per second the envelope of the received signal crosses a level r in a given sense is denoted by $N(r)$. Rice[3] has predicted that $N(r)$ will be directly proportional to the probability density function of the amplitude of the signal at the receiver. The measurements described in Reference 1 show that for this link the amplitude distribution corresponds, under normal conditions, to the sum of a constant vector plus a Rayleigh distributed vector; $N(r)$ should therefore be of the form

$$N(r) = A\frac{2r}{k_0^2}\mathrm{I}_0\left(\frac{2r}{k_0^2}\right)\exp\left[-(1+r^2)/k_0^2\right]$$

where A is a constant and the amplitude r and the r.m.s. value k_0 of the Rayleigh component are measured relative to the amplitude of the constant vector and I_0 is the modified Bessel function of order zero. The limiting form as the constant vector tends to zero will, of course, be the simple Rayleigh distribution. It follows, therefore, that, if the quantity

$$\phi(r) = \int_0^r N(r)dr$$

is expressed as a percentage of $\phi(\infty)$ and plotted on Rayleigh graph paper against the level r in decibels, the curve should approximate fairly closely to a straight line. The slope should be unity for a pure Rayleigh signal and somewhat larger if a background signal is present.

The function $N(r)$ was determined experimentally for about 30 different periods, and in each case the curve could be represented reasonably well by an expression of the form given above. The function $\phi(r)$ was obtained by numerical integration of the $N(r)$ curve. The results of six experiments are shown plotted on Rayleigh paper in Fig. 14. All the curves can be represented reasonably well by straight lines of slopes near unity, in agreement with the prediction of Rice.

(7) POWER SPECTRUM AND AUTO-CORRELATION MEASUREMENTS AT S-BAND

The detected signal output from the receiver consists of a randomly varying a.c. component fluctuating about a mean d.c. level, the power in the a.c. component being distributed co tinuously throughout the whole of the frequency spectru The median-level fading rate discussed in the previous Secti presumably indicates the frequency about which a large propo tion of the a.c. power is concentrated but gives no indication the upper frequency limit beyond which the energy content c be regarded as negligible. A knowledge of this is clea desirable since it determines the frequency response that recor ing apparatus must have if an accurate picture of the sign variations is to be obtained.

The average power dissipated in the frequency range $(0, g)$ the detected signal $r(t)$ flowing into a 1-ohm resistor can written

$$P(g) = \overline{r}^2 + \int_0^g \omega_{ac}(f)df$$

where $\omega_{ac}(f)$ is the power spectrum of the a.c. component the signal. $\omega_{ac}(f)$ can be most easily evaluated in terms of t closely related auto-correlation function $\psi_{ac}(\tau)$ from the relati

$$\omega_{ac}(f) = 4\int_0^\infty \psi_{ac}(\tau)\cos(2\pi f\tau)d\tau$$

where $\psi_{ac}(\tau)$ is defined as

$$\psi_{ac}(\tau) = \frac{1}{T}\int_0^T [r(t) - \bar{r}][r(t+\tau) - \bar{r}]dt$$

T being the interval over which the signal is studied.

The experimental procedure consisted in estimating the fun tion $\psi_{ac}(\tau)$ by numerical integration of a high-speed pen reco of the signal over a time T and then taking its Fourier cosi transform, again numerically, to obtain $\omega_{ac}(f)$. Since t propagation conditions can only be regarded as stationary ov a fairly short period, different results will be obtained from ho to hour. This is evident from the results reported in the previo Section on the variation in the median-level fading rate throug out a day. For complete information about the power spectru experiments should be performed over a wide variety of co ditions. The amount of numerical work involved makes th impractical and here only one period of 10 sec, which appear to be representative of the more usual type of signal variati and which had a median-level fading rate of 2 c/s, was studi

A plot of $\psi_{ac}(\tau)$, normalized to $\psi_{ac}(0)$, for the 10 sec peri

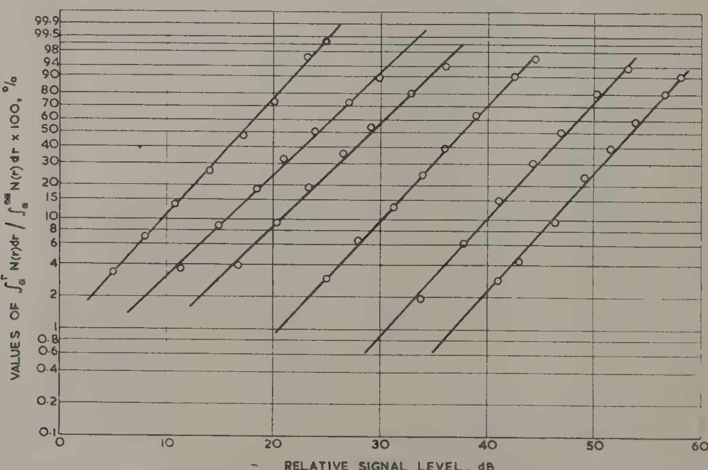
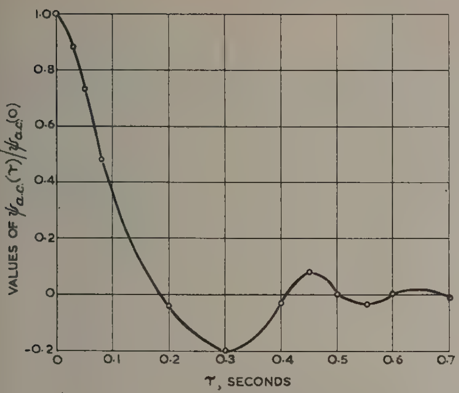
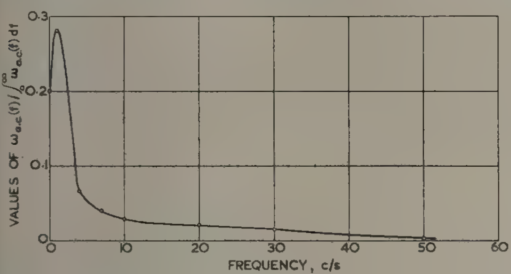


Fig. 14.—Distribution of the integral of the S-band fading rate with level for the Start Point–Wembley path.



ig. 15.—Normalized a.c. auto-correlation function of the S-band detected signal for the Start Point–Wembley path.

xamined is shown in Fig. 15. The general form is similar to at of a uniform distribution of power in a narrow band, but, nlike the latter case, the zeros of $\psi(\tau)$ are not equally spaced. he a.c. part of the power spectrum derived from this and ormalized to the total a.c. power in the wave is shown in Fig. 16. s expected from median-level fading rate, the a.c. power density



ig. 16.—Normalized a.c. power spectrum of the S-band detected signal for the Start Point–Wembley path.

or this fairly typical signal has a maximum value in the neighbourhood of 2 c/s. When the calculated value of $\bar{r}$ is taken into ccount it is found that about 75% of the power is d.c. and that nore than 99% of the total power lies in the frequency range –50 c/s.

(8) DISTRIBUTION OF FADES AT S-BAND

A tropospheric link, under unfavourable weather conditions, nay be operating with the median value of the signal only a ew decibels above the noise level of the receiver. Under such onditions excessively long fades at levels just below the median vould severely restrict the usefulness of the system. It is clear, herefore, that an evaluation of the reliability of a communication ystem in which the signal fluctuates over a wide range should ake into account the distribution of the duration of fades as a unction of level. A limited amount of work on this problem as been undertaken for the S-band Start Point–Wembley link.

The duration of fades has been estimated both from high-peed pen records and electronically. The electronic equipment ised a continuously-variable gating circuit in conjunction with a Dekatron counter, and was arranged so that only fades of greater luration than the 'gate shut' time were counted. By gradually educing the 'gate shut' time an analysis of the distribution of he duration of fades could be made. The disadvantge of this nethod was that the complete distribution at a given level took ıp to 45 min to obtain, and it is doubtful whether the conditions remained sufficiently stable over such a period to give very accurate results. Far more weight was given to the results obtained from high-speed pen records.

(8.1) Average Duration of Fades at Different Levels

The average duration of fades below a given level r can be written

$$\bar{t}(r) = \frac{Pr[r(t) < r]}{N(r)}$$

where $N(r)$ is the number of times per second the detected signal crosses a level r with a negative slope. Rice[4] has shown that, for a Rayleigh distributed amplitude,

$$N(r) = \left(\frac{2B}{\pi}\right)^{1/2} \frac{r}{k^2} \varepsilon^{-r^2/k^2}$$

where

$$B = 4\pi^2 \int_0^\infty \omega(f)(f - f_0)^2 df$$

$\omega(f)$ is the r.f. power spectrum and f_0 is the frequency of the transmitter. In this case, therefore,

$$\bar{t}(r) = \left(\frac{\pi}{2B}\right)^{1/2} \frac{k^2}{r}(\varepsilon^{r^2/k^2} - 1)$$

If $\bar{t}_0$ is the average fading duration at the median level and ξ denotes the level r measured relative to the median value, it is easy to show that

$$\frac{\bar{t}(r)}{\bar{t}_0} = \frac{\varepsilon^{0 \cdot 693\xi^2} - 1}{\xi}$$

It should be noted that $\bar{t}/\bar{t}_0$ is independent of the power spectrum and depends only on the assumed form of the amplitude distribution. A plot of the above equation together with two sets of experimental results, one obtained from a pen record and the other by the electronic timer, are shown in Fig. 17. Both sets of results are distributed about the theoretical curve, verifying that the signal-amplitude distribution is of the Rayleigh type. As expected, the results obtained from the electronic timer show more spread than those estimated from the pen records.

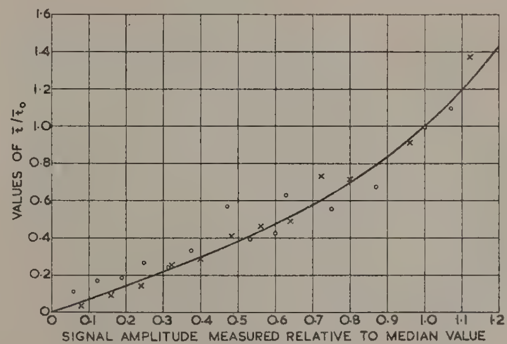


Fig. 17.—Average duration of fades at S-band as a function of level for the Start Point–Wembley path.

○ Results from electronic counter.
× Results from a high-speed chart.

(8.2) Distribution of the Duration of Fades

A theoretical discussion of the distribution of the duration of fades as a function of the level has recently been given in Reference 4. In this paper graphs of a function $F_r(u, r)$, defined as the probability that when a signal $r(t)$ crosses downwards through a level r it will remain below that level for more than τ seconds, are given for various values of r for the particular

case of a Rayleigh-distributed amplitude and a Gaussian power spectrum

$$\omega(f) = \frac{1}{\sqrt{(2\pi)}\sigma} \exp\left[-(f-f_0)^2/2\sigma^2\right]$$

The variable u is defined as $\tau/\bar{t}(r)$, and from the previous discussion it is clearly a function of the parameter σ defining the power spectrum. The quantities $F_r(u, r)$ and u can both be determined experimentally. A comparison of the practical curves with those of Rice will give an indication of whether the Gaussian form of the power spectrum, often quite arbitrarily assumed, is a valid approximation.

Fade distributions at three levels, determined from a 35 sec sample of high-speed chart, together with the theoretical curves of Rice, are shown in Fig. 18. The record used showed no

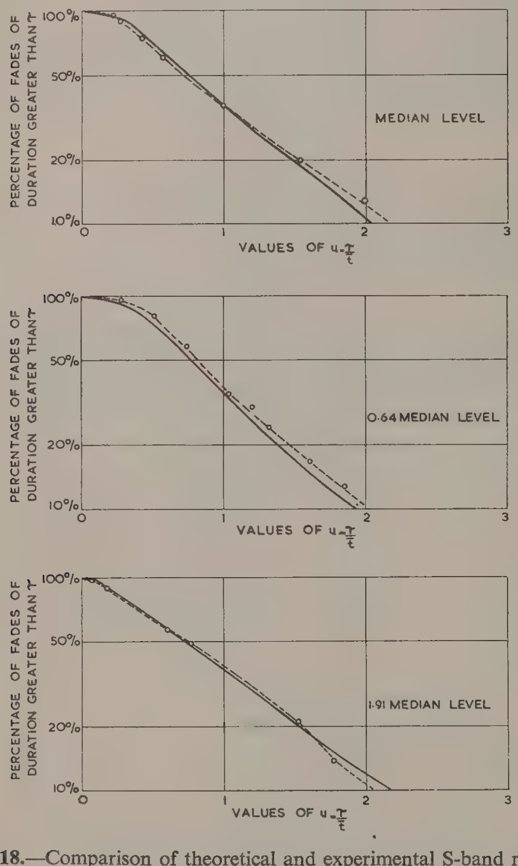


Fig. 18.—Comparison of theoretical and experimental S-band results for the distribution of duration of fades below a given level for the Start Point–Wembley path.

——— Theoretical.
○— —○ Experimental.

unusual features, and the fading rate at the median level was about 3 c/s. There is seen to be quite close agreement between the practical and theoretical results. The distributions obtained from the electronic timer, although of the same general form, did not follow the theoretical curves nearly as closely, possibly because of a variation in $\bar{t}$ over the time necessary to obtain a complete distribution at a given level. Clearly, before any general conclusions can be drawn, many more results obtained under a variety of conditions would have to be studied. However, the few measurements made suggest that, under normal conditions at least, the assumption of a Gaussian power spectru is probably valid.

(9) X-BAND MEDIAN LEVEL MEASUREMENTS AND CO PARISON WITH S-BAND RESULTS AT WEMBLEY

The pulsed X-band scatter system between Start Point a Wembley was put into operation in June, 1958, and propagati data, extending over the period of a year and collected at t rate of a few days recording during each month, has be analysed. The number of hours of recording during each mon is shown in Table 5.

Table 5

Number of Hours during which Data were Recorde

Month, 1958	June	July	August	Sept.	Oct.	Nov.	De
Number of hours of recording	155	112	36	239	154	116	10
Month, 1959	Jan.	Feb.	March	April	May	Total in ye	
Number of hours of recording	181	121	96	30	96	1 436	

It is seen that, with the exception of August, 1958, and Apr 1959, when certain modifications to the equipment were mac the link operated for four or more days during each mont Sufficient data should therefore be available to give a reasonab accurate picture of the signal variation throughout the year. at S-band, the performance of the link was measured by t hourly median value of the system transmission loss, which w estimated by eye from the pen record of a low-speed recorde At S-band, the difference between the hourly median leve computed accurately from the counter readings of the sign sampling machine and those estimated by eye from a low-spee chart was usually less than 1 dB. The estimates of the X-bar median level may therefore be expected to have an accuracy this order or somewhat better. Whenever possible, the X- an S-band links were operated simultaneously. In addition providing a general picture of the propagation at X-bar throughout the year, information was obtained about the fr quency dependence of the transmission characteristics of scatter propagation path over quite a wide range of atmospher conditions. Further, since aerials of the same size were use in the S- and X-band equipments, a direct comparison of th system transmission loss at the two frequencies will provide basis for deciding whether, for equipments of comparable bu one frequency has a decisive advantage over the other.

A summary of the variations in the hourly median values the transmission loss at X- and S-band for corresponding perioc of 24 hours and within a month are given in Table 6. It is see that the diurnal variations at X-band were comparable to thos at S-band, the difference at the two frequencies being typicall 3–4 dB. The averages of the diurnal variations of all th measurements shown in Table 6 are, for both frequencies, abou 14·5 dB, while the average values for the month are respectivel 25·6 dB and 30·5 dB at X- and S-band. The measured extreme of hourly median transmission loss at X-band were 118 an 172 dB, the range of 53·5 dB being of the same order as th yearly S-band figure (60 dB). The mean variations in hourl medians within periods of 6 hours at X-band are shown i Table 7.

These values are only slightly greater than the correspondin S-band results and indicate that the X-band signal is, on th whole, equally variable throughout the day.

A variation in hourly median level of more than 20 dB i 24 hours was usually an indication of the occurrence of a ducted

Diurnal Variations in Median Level

Test period	1 X	1 S	2 X	2 S	3 X	3 S	4 X	4 S	5 X	5 S	6 X	6 S	7 X	7 S	8 X	8 S	9 X	9 S	10 X	10 S	11 X	11 S	Total monthly variation X	Total monthly variation S
1958																								
June	10·1	5·3	15·0	10·0	11·0	9·2	11·3	—	14·5	—	12·8	—											32·7	29·0
July	20·9	30·6	12·7	11·2	30·4	43·1	12·4	7·1	8·4	8·4													32·5	46·9
August ..	22·1	20·0																					22·1	20·0
September ..	12·0	10·2	21·0	24·7	10·3	8·0	10·0	7·0	10·9	8·9	11·8	—	7·5	7·2	9·7	11·7	12·2	7·6	12·4	7·7	17·9	15·0	30·0	42·5
October ..	14·5	10·7	15·0	13·7	11·1	10·5	18·0	11·2	8·1	10·6	29·1	35·9	16·2	—									36·5	49·4
November ..	18·0	17·9	18·0	16·8	13·5	11·2	13·4	9·3	8·6	6·8													18·6	29·3
December ..	18·5	9·0	12·3	—	14·7	16·8	19·8	—															23·5	20·2
1959																								
January ..	16·2	13·8	16·3	10·0	14·6	11·4	27·7	46·5	29·7	25·0	22·3	38·5	10·8	15·2									35·0	53·0
February ..	16·1	—	11·0	8·6	10·6	9·5	9·1	7·8															19·0	23·8
March ..	11·2	10·7	7·2	8·9	12·8	12·5																	16·9	19·5
April	14·0	13·5																					14·0	13·5
May	16·6	10·0	9·1	15·1	19·4	15·2																	26·7	19·2

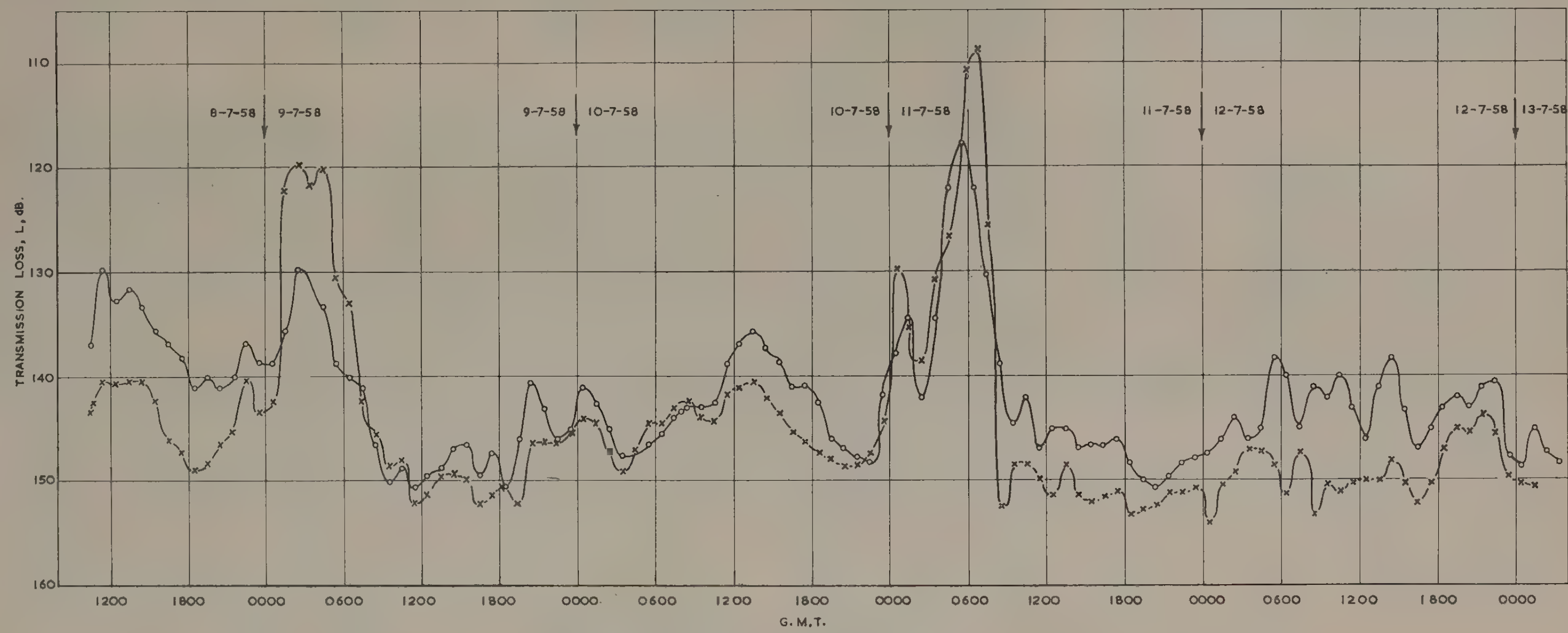


Fig. 19.—Comparison of the X- and S-band median-level transmission loss for the Start Point–Wembley path for July, 1958.
×—×—×— S-band. ○—○—○ X-band.

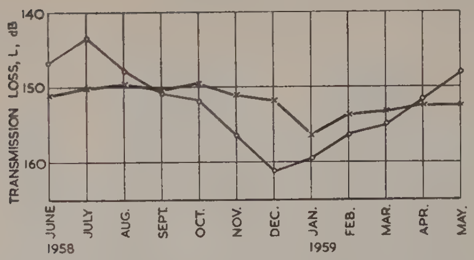


Fig. 20.—Comparison of the monthly median transmission loss at S- and X-band for the Start Point–Wembley path.

○——○ X-band.
×——× S-band.

Table 7

Period, hours	0000–0600	0600–1200	1200–1800	1800–2400
Average variation, dB	7·5	8·4	7·5	8·3

type signal, already described in connection with the S-band results, which persisted for an hour or longer. During the course of the year such signals were observed at X-band on eight occasions, and reference to Table 6 will show that in each case a similar signal was also present at S-band. The time at which the high-level signal appeared and its duration were almost identical at the two frequencies, although the increase in level was usually 5–12 dB greater at S- than X-band. A plot of measured values of the hourly median transmission loss at X- and S-band, which includes two periods of ducting, is shown in Fig. 19 for a 4-day period of continuous recording. The similarity of the variations shows that the slow fading at the median level cannot be effectively dealt with on a frequency-diversity basis. As might be expected from the similarity in median-level variation at X- and S-band, the X-band signal showed no well-marked diurnal variation.

The distribution of the X-band hourly medians within a month in which four or more days recording were possible was found to be approximately Gaussian. The average value of the monthly standard deviation was about 5 dB, very nearly the same as the corresponding S-band figure. Plots of the monthly median values of the hourly medians for S- and X-band for the year June, 1958 to May, 1959 are given in Fig. 20. It is seen that the seasonal dependence of the signal level is far greater at X- than S-band. The range of monthly medians is 17 dB at X-band compared with 7 dB at S-band, the X-band transmission loss being a few decibels less than that for the S-band during the summer, but, on the average, about 5 dB greater during the winter months.

The yearly distribution of the X-band hourly medians, together with the distribution for the best (July) and worst (December) months are plotted in Fig. 21. For the two months (August, 1958 and April, 1959) when less than two days' recording were obtained, monthly median values were estimated and normal distributions assumed with standard deviations equal to the mean value for the other ten months. The X-band distribution for the year has a median value of about 152·5 dB and a standard deviation of 7·3 dB; these figures should be compared with the typical S-band values of 151 and 5·5 dB. Reasonable estimates of the aerial coupling loss at S- and X-band are respectively 3 and 8 dB. Using these figures, it is interesting to note that the formula of Reference 2 predicts a median signal level difference between the X- and S-band systems of about 4 dB. This is in good agreement with the measured difference between the yearly

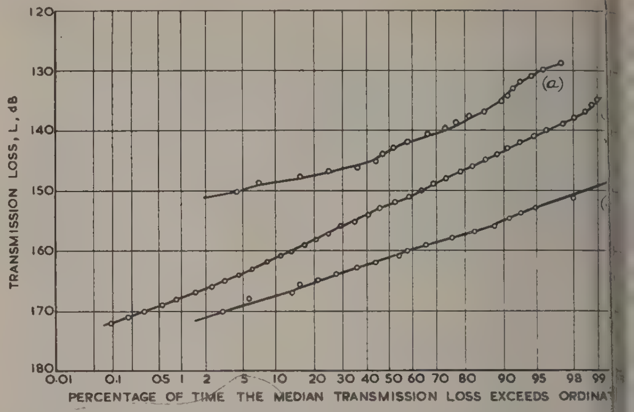


Fig. 21.—Yearly distribution of the X-band hourly median transmission loss for the Start Point–Wembley path.

(a) Best month (July).
(b) Year.
(c) Worst month (December).

median values. However, the 6 dB discrepancy between the predicted and measured values of the S-band transmission loss at Witnesham serves to emphasize that the theoretical descriptions of scatter propagation, while being extremely useful guides to the performance of a proposed propagation link, may, in particular cases, be in error by up to 6 dB. In practice this could be quite serious, since it might, for example, mean the difference between a 10 ft and a 20 ft paraboloid.

(10) COMPARISON OF X- AND S-BAND MEDIAN-LEVEL FADING RATES

Over the period June, 1958 to May, 1959 about 140 simultaneous measurements of the X- and S-band median-level fading rates were made, a sampling period of 2–3 min being used. In addition to spot measurements throughout the year, estimates of the fading rates at regular intervals throughout the period 0900–1700 hours were made on several occasions.

Cumulative distribution plots of the measured fading rates at X- and S-band are shown in Fig. 22; the fading rates have been normalized to the median values which at S- and X-band

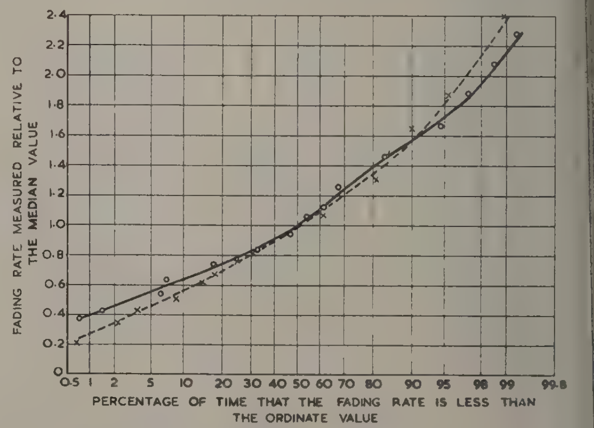
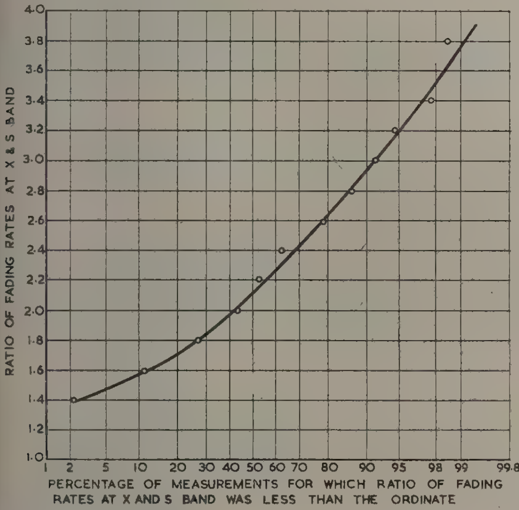


Fig. 22.—Normalized distributions of X- and S-band fading rates measured at Wembley.

○——○ X-band.
×——× S-band.
Median fading rate at X-band: 4·8 c/s.
Median fading rate at S-band: 2·4 c/s.

e, respectively, 2·4 and 4·8 c/s. The distributions at the two :quencies are seen to be quite similar, with about 90% of the easurements in each case lying within the range 0·4 to 2·0 times e respective median values.

The distribution of the ratio of the simultaneous fading rates the two frequencies is of considerable interest. Theoretical scriptions[5] of the fading of a tropospheric signal have predicted at, for a given propagation path, the median-level fading rates ll be proportional to f^{γ} where γ lies between two-thirds and ity, the former value being associated with the random self otion of the scatterers and the latter with their steady drift locity. In the present case the ratios of the X- and S-band easurements would be expected to lie in the range 1·97 and 77. A cumulative distribution plot of measured values of this tio is shown in Fig. 23. It is seen from the Figure that the



g. 23.—Distribution of the ratio of fading rates at X- and S-band measured at Wembley.

lues of 1·97 and 2·77 correspond, respectively, to the 41% d 84% levels, so that only 43% of the measurements lie within e expected limits. It is of interest to note, however, that the tio of the median values of the fading rates at X- and S-band 2·0, corresponding almost exactly to the theoretical lower nit. As might be expected from the large spread in the ratio the fading rates at the two frequencies, there was little etailed correlation between the variations in fading rates at - and S-band throughout the day. A typical result, showing pairs of measurements taken over a period of about 3½ hours given in Fig. 24. All that can be said is that the general ends over this period are similar.

(11) AERIAL COUPLING LOSS AT X-BAND

In this Section a series of measurements to investigate the rial coupling loss of the X-band system are described. At e receiver terminal a pyramidal horn of 28·7 dB gain was ounted on the tower immediately above the X-band paraoloid, and a waveguide switch was incorporated in the system allow the receiver to be switched rapidly from one aerial to e other. The signal was received in alternate 3 min periods the paraboloid and horn, the level being continuously recorded the low-speed chart. The change in median level, ΔL, as e receiver was switched from the paraboloid to the horn was

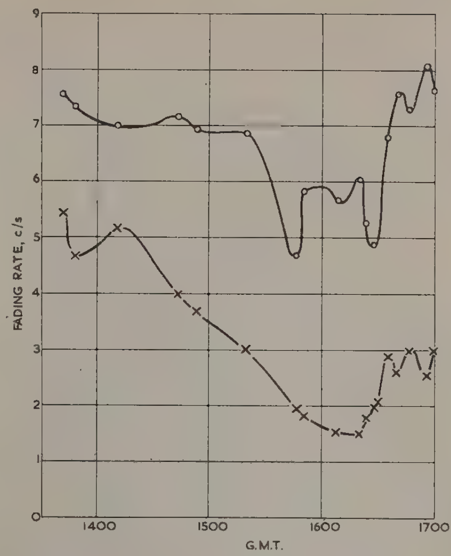


Fig. 24.—Variation of the X- and S-band fading rates during a 3-hour period measured at Wembley.

O——O X-band.
×——× S-band.

estimated from the chart. From this and the known difference, ΔG, between the plane-wave gains of the two aerials, the difference in aerial–median coupling loss, ΔL_c, for the system using two paraboloids and the system using a paraboloid at the transmitter and the horn at the receiver, was calculated from the formula

$$\Delta L_c = \Delta G - \Delta L$$

In all, 95 estimates of ΔL were made, the measurements extending throughout March, 1959. The spread in the computed values of ΔL_c, ranging from almost zero up to a maximum of 12 dB, is shown in the histogram of Fig. 25. This distribution is very approximately Gaussian with a median value of 4·4 dB and standard deviation of about 2·3 dB. Fig. 26 shows values of ΔL_c obtained in 13 consecutive periods of 6 min. The wide

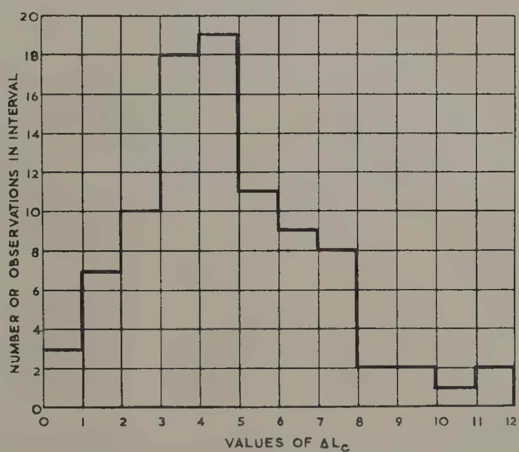


Fig. 25.—Histogram of the measured difference, ΔL_c, in aerial coupling loss at 9·640 Gc/s for a system using two 8 ft paraboloids and a system using one paraboloid and a horn of gain 28·7 dB for the Start Point–Wembley path.

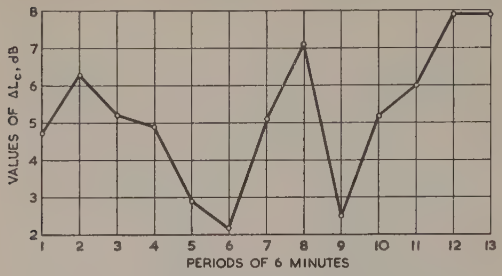


Fig. 26.—Variation of ΔL_c with time of day.

fluctuation in the estimated values of ΔL_c from one period to the next was typical of all the measurements made.

Although ΔL_c is a convenient parameter to measure and brings out the time-dependent nature of the coupling loss, the quantity of greater practical interest is the coupling loss of the system when paraboloids are used at both terminals. Theoretical estimates of ΔL_c can be made from References 6 and 7, but the average of these is 2·0 dB—nearly 2·5 dB less than the measured median value of 4·4 dB. A possible explanation of the discrepancy may be that the effect of ground reflections at the receiver terminal introduced a systematic error into the measurements which led to an overestimate of ΔL_c. In view of this it is difficult to deduce from the measurements the coupling loss to be expected from the two-paraboloid case. The figure of 8·0 dB used in Section 9 was reached by averaging the various theoretical estimates given in the literature.

The difficulty arises because the measurements determine a small difference between two coupling losses, and even a relatively small error can be of considerable significance. The results emphasize the advisability, whenever transmitter power permits, of arranging that both the receiver and transmitter can be switched simultaneously to low-gain aerials so that the coupling loss of the original system can be measured directly.

(12) CONCLUSION

An S-band signal transmitted from Start Point, using a 500-watt c.w. magnetron, was monitored simultaneously at Wembley and Witnesham by identical receivers, the distances from the transmitter in the two cases being 173 and 247 miles. The median values of the system transmission loss at the two sites over the period of about a year were, respectively, about 151 and 162 dB. In each case the hourly median values of the transmission loss over a month or longer were approximately normally distributed with a standard deviation of about 5 dB. The variations in median level at the two sites usually showed the same general trends but there was little correlation in the detailed fluctuations. However, measurements were made when the median signal at Witnesham was in some cases higher and in other cases as much as 30 dB lower than that at Wembley. The median difference in hourly median levels at the sites was 11·6 dB, corresponding to an average attenuation rate of the S-band scatter signal of 0·16 dB/mile, in agreement with other reported measurements. The transmission was also monitored by the D.S.I.R. at Slough, about 13 miles from Wembley; the median levels measured simultaneously at these two stations were, in general, quite highly correlated. Experiments comparing the signal received over the same path at good and bad sites showed that the average difference in level could be estimated approximately from the angular path distances corresponding to the two sites.

Simultaneous measurements of S- and X-band transmissi[ons] over the Start Point–Wembley path were made for the perio[d of] a year. The variations in hourly median values of the sys[tem] transmission loss were, on the whole, very similar at the t[wo] frequencies; periods of ducting observed at S-band were du[pli]cated, both in time of occurrence and duration, almost exa[ctly] at X-band, although the increase in level under such conditi[ons] was usually somewhat greater at S- than X-band. A seaso[nal] dependence, the signal being higher in the summer than [in] winter, was observed at both frequencies. The X-band dep[en]dence was, however, greater than at S-band, the yearly rang[e of] monthly median values of signal level being 17 dB at X-ba[nd] compared with an average of 7 dB at S-band. The yea[rly] median value of the system transmission loss at X-band w[as] about 153 dB.

The ratio of the X- to S-band fading rate at the median sig[nal] level was found to vary over quite a wide range, less than 5[%] of the measured values falling within the predicted limits.

At S-band, the fading rate was studied as a function of le[vel] and was found to have the same distribution as the amplit[ude] of the detected signal. A limited amount of work on [the] distribution of the duration of fades below a given level sugge[sts] that the frequently made assumption, that the r.f. power spectr[um] of an incoming scatter signal is Gaussian, is probably valid un[der] normal propagation conditions.

An auto-correlation analysis of a fairly typical sample o[f a] high-speed pen record of the S-band signal showed that, un[der] normal conditions, 99% of the total energy of the detected sig[nal] was contained in the frequency range 0–50 c/s.

Measurements at X-band indicated that the coupling loss [of] an aerial system varies with time over a wide range.

(13) ACKNOWLEDGMENTS

The work described was sponsored jointly by the Ministry [of] Supply and the Admiralty, to whom thanks are due for perm[is]sion to publish the paper.

We are indebted to the Director of Radio Research D.S.I.[R.] for permission to publish the observations made at Slough us[ed] in the paper.

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PARABOLOIDAL REFLECTORS WITH AXIAL EXCITATION

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SUMMARY

The reflectors are deep paraboloids having axially oriented dipoles the foci. Odd-function aperture distributions and radiation patterns derived by approximate methods and are compared with an perimental polar diagram. In spite of a null along the axis, strong upling at short range is observed between two aligned aerials. This associated with a coupling diagram of abnormal directivity, which y be explained by a principle of field-fitting between the incident d and the normal aperture distribution. A reduction of Maxwell's uations in paraboloidal co-ordinates to Bessel's equation is appended.

LIST OF SYMBOLS

$4a$ = Latus rectum.
$F(y)$ = Aperture distribution; variation of electric field intensity.
c_n = nth component in Fourier series.
θ = Polar co-ordinate measured from dipole axis.
λ = Wavelength.
u = $(\sin\theta)/\lambda$.
b = Aperture width.
$G(u)$ = Fourier transform of $F(y)$.
T = A displacement on the y-axis.
(T) = Convolution of $F(y)$ and $G(y)$.
P_T = Powers received and transmitted.
v, ϕ = Paraboloidal co-ordinates.
l = A scaling factor.
ϕ, z = Polar co-ordinates.
G = Gain and an intermediate vector.
ω = Angular frequency.
μ_0 = Permittivity and permeability of free space.

(1) INTRODUCTION

It is customary for the dipole in a paraboloidal reflector to be aced at the focus and perpendicular to the axis of revolution. e reflecting surface does not extend beyond the focal plane. it did, radiation by paths including reflections from certain eas of the inner and outer parts of the surface would arrive a distant point in antiphase. Looking into the aperture, the nduction currents would be equivalent to a dipolar current stribution in the directrix plane.

If the dipole is perpendicular to the axis of revolution, two the directions of strongest primary radiation are trouble-me. Primary radiation towards the apex creates a standing ve which is such that, during a test with axial motion of the pole, the geometrical focus may be masked.[1] In the opposite ection is non-reflected primary radiation. To prevent its erference with intended reflections from the paraboloid, a w obstacle is often introduced in the centre of the aperture at is being illuminated. A dilemma then arises regarding the size of the obstacle, for it obstructs reflections from the central regions of the paraboloid as effectively as it shrouds the dipole.

These problems are transformed when the dipole is put along the axis of revolution. The directions of zero primary radiation are just those which were formerly troublesome. There is no change of phase in reflections as the focal plane is crossed, and the paraboloid need not therefore be shallow. The dipolar distribution is exchanged for a symmetrical radial pattern resembling the cross-section of an E_0 waveguide mode. The symmetry holds promise of a simplified solution to Maxwell's equations. These attractive properties of axial excitation are offset by the realization that the radiated field along the axis is necessarily zero, since the aperture distribution is an odd function. This means that only hollow beams of radiation are possible.

Experimental results for parabolic reflectors with transverse dipoles are plentiful. Those given by Moullin[1] are accompanied by a critical discussion of the employment of parabolic surfaces at non-optical frequencies.

The solution of Maxwell's equations inside a deep paraboloid of revolution is performed by Pinney[2,3] in terms of Laguerre polynomials. Another solution in the form of complex integrals is given by Shalshaya[4] for the case of a transverse dipole. In the Appendix of the present paper, it is shown that a solution could be expressed in terms of Bessel functions for the E_0 configuration.

Another object of the paper is to consider the coupling between two identical paraboloids with axial excitation. This may take the form of a coupling between main lobes, requiring each paraboloid axis to be offset from the line joining the aerials. Such a coupling reveals the normal polar diagram when one of the aerials is turned. A more interesting form of coupling results from alignment of the paraboloids on a common axis at short range. The radial field pattern transmitted by one aerial then fits the similar pattern demanded by the other, as if there were two cartwheels sharing a common hub. Strong coupling is to be expected.[5] If one of the aerials is now turned, a rapid weakening of the coupling occurs, since the patterns fit badly together, like cartwheels with eccentric hubs. Questions arise concerning the prediction of the modified polar diagrams, the possible achievement of enhanced directivity without super-gain technique, and the efficient reconcentration of energy at the second focus.

(2) APPROXIMATE THEORY FOR THE RADIATION PATTERNS

(2.1) Estimation of the Aperture Distribution

If the electric and magnetic field strengths were known over an aperture plane enclosing the mouth of the paraboloid, the radiation pattern could be found. A first attempt at the aperture distribution is made in terms of geometrical optics (Fig. 1).

We express the equation to a cross-section of the paraboloid in rectangular co-ordinates

$$y^2 = 4ax \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

Written contributions on papers published without being read at meetings are ited for consideration with a view to publication.
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Mr. Donaldson is now with Associated Electrical Industries (Rugby), Ltd., Mr. nch is with the Radio Corporation of America, and Dr. Midgley is Reader in ctrical Engineering at the Bradford Institute of Technology.

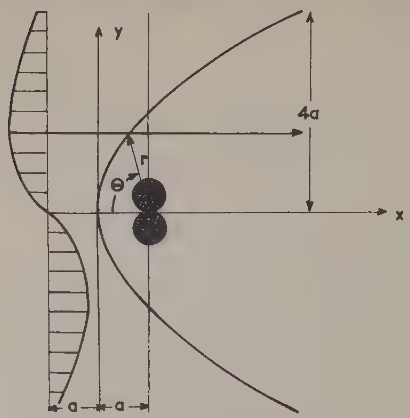


Fig. 1.—The parabolic profile.

(a) The black area represents the primary polar diagram of the dipole.
(b) The lightly hatched area is the aperture distribution derived from (a) by projection on to the directrix plane.

The primary radiated field strength from the dipole is approximately proportional to $(\sin\theta)/r$, where θ is measured from the axis, and is related to the rectangular co-ordinates by the equation

$$\tan\theta = \frac{y}{a - x} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

Combining eqns. (1) and (2) and solving in terms of y and $\sin\theta$, we have

$$F(y) = \frac{4ay}{4a^2 + y^2}\left[\left(\frac{y^2}{4a} - a\right)^2 + y^2\right]^{-1/2} \quad . \quad . \quad (3)$$

$F(y)$ is the variation in field strength across the aperture plane on the supposition that rays travel along a radius from the dipole to a point on the parabolic surface and are reflected parallel to the axis, total path lengths being the same for all rays which emerge co-phased in the aperture plane. $F(y)$ is taken to be zero outside the radius of the finite aperture, and the contribution from non-reflected primary radiation is ignored.

(2.2) Fourier Transforms

The transform of eqn. (3) would lead to the radiation pattern, but the integral does not appear to exist in closed form. In the absence of the direct transform, $F(y)$ is analysed into sinusoidal components numerically for the particular case where the aperture radius is equal to $4a$. Only odd sine terms occur and these are given by the equation

$$c_n = \frac{1}{\pi}\int_{-4a}^{4a}\frac{4ay}{4a^2 + y^2}\left(\sin\frac{n\pi y}{4a}\right)\left[\left(\frac{y^2}{4a} - a\right)^2 + y^2\right]^{-1/2} dy \, . \quad (4)$$

The first, third and fifth components are in ratios 100 : 17 : 4. The transforms of the separate components are given by the following integral, using b for aperture width and $u = (\sin\theta)/\lambda$:

$$G(u) = \int_{-b/2}^{b/2}\sin\frac{2\pi ny}{b}\,\varepsilon^{-j2\pi uy}dy \quad . \quad . \quad . \quad . \quad (5)$$

$$= \frac{bn\pi\sin\pi bu}{(n\pi)^2 - (\pi bu)^2} \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

The notation follows that of Ramsay,[6] and eqn. (6) extends his results. Curves illustrating the first, third and fifth components are shown in Fig. 2. They are odd functions with two

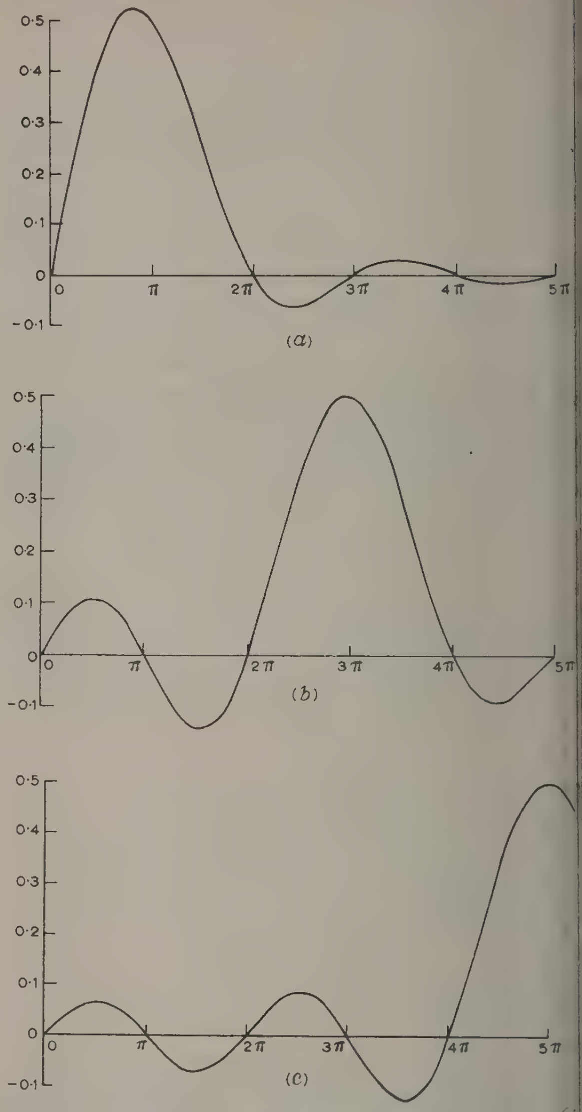


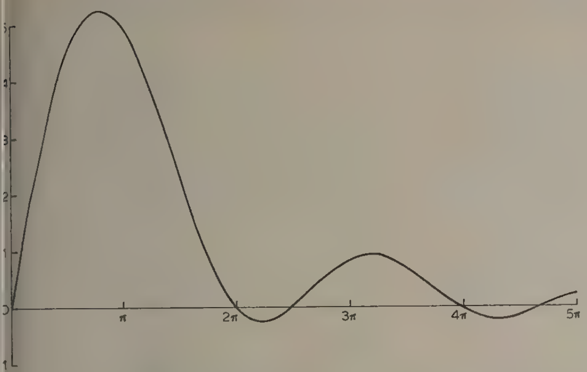
Fig. 2.—Fourier transforms of $(\sin 2\pi ny)/b \quad -\tfrac{1}{2}b < y < \tfrac{1}{2}b$.
Vertical axes, $G(u)/b$.
Horizontal axes, πbu.
(a) $n = 1$; (b) $n = 3$; (c) $n = 5$.

main lobes, which move further from the origin the higher t[he] value of n, in which respect they repeat the behaviour of th[e] even counterparts.[5]

An approximate resultant transform appears in Fig. 3 aft[er] numerical combination of these first three terms of the seri[es]. Further continuation of the series is hardly justified in view [of] the geometrical-optical approximations in forming the apertu[re] distribution.

(2.3) Convolution of the Fourier Transform with the Apertu[re] Distribution

The coupling that is achieved between two aerials is believ[ed] to be dependent upon the product of the field the receiving aer[ial] actually encounters and the field it can best use.[5, 8] The latt[er] has the form of the aperture distribution during transmissio[n]



. 3.—Fourier transform of the approximate aperture distribution.

:ept that it is a travelling wave in the opposite direction. Let y) of eqn. (3) be taken as the desired, or 'complementary', ld in the aperture plane of the receiving aerial.

Given an aperture width exceeding 10λ it is usual to accept Fourier transform as the radiation pattern, since cosine :tors vary relatively slowly. Further, if θ is small enough to replaced by $\sin\theta$, the incident field at finite range may be presented as the Fourier transform with y as the independent riable. For,

$$u = (\sin\theta)/\lambda \simeq \theta/\lambda \simeq y/r\lambda$$

d we may write

$$G(u) = G(y/r\lambda) = G'(y)$$

Let the transmitting aerial now be turned through an angle θ. ig. 4). The distribution incident upon the receiver is shifted

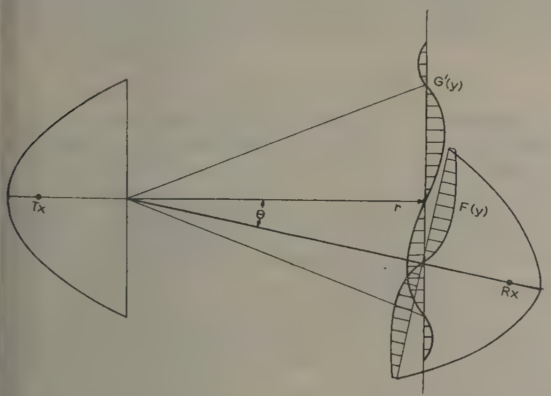


Fig. 4.—Misaligned paraboloids to illustrate the convolution of aperture distribution and radiation pattern.

a displacement, $T \simeq r\theta$, on a scale of y with origin fixed at the ntre of the receiver aperture. Taking the products of the two stributions and integrating with respect to y over the aperture dth for a series of displacements T, we generate the convolution egral $\Phi(T)$.

$$\Phi(T) = \int_{-b/2}^{b/2} F(y)G'(y - T)dy \quad . \quad . \quad . \quad . \quad (7)$$

$\Phi(T)$ represents the variation in power received and might be med a coupling diagram as distinct from a polar diagram. The upling diagram is a function of range and depends upon the lar diagrams of both aerials, which participate in the coupling.

$\Phi(T)$ is not intended to give the absolute power transferred. For this purpose a scaling factor would have to be included and the absolute level of $F(y)$ would be required, which would itself require knowledge of the power transferred. Eqn. (7) is also oversimplified in its one-dimensional form. As it stands it applies to a related problem of parabolic cylinders, where nothing varies along the third co-ordinate axis. In the paraboloid-of-revolution problem, the equation represents conditions over a thin slice across the aperture, taken along the line joining the hubs of the cartwheel picture.

A numerically evaluated integral is illustrated in Fig. 5(*b*). This

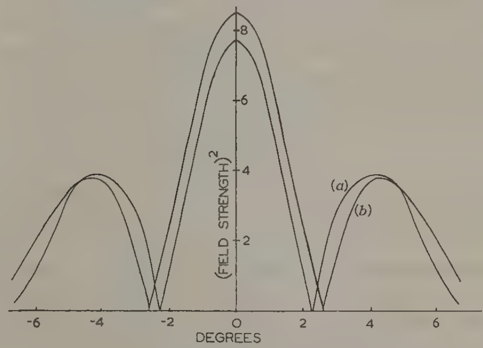


Fig. 5.—Computed coupling diagrams.

(*a*) Convolution of one cycle of sine wave with its Fourier transform.
(*b*) Convolution of the approximate aperture distribution with its Fourier transform.

is a convolution of $F(y)$ from eqn. (3) and $G'(y)$ from Fig. 3. A range of 5 m is chosen, since it causes the first zero in $G'(y)$ to occur at the edge of the aperture, when there is no displacement. Thirty-two ordinates are taken of each function over the aperture plane; these are paired according to the particular displacement and are then multiplied with summation of products to yield one point on the coupling diagram.

The coupling diagram is an even function with principal maximum at the origin, whereas both of the functions from which it is derived are odd and zero at the origin.

(2.4) Attenuation with Range

In the common-axis arrangement there is what might be described as a critical range at 5 m, which is the greatest separation for which the main lobes of the transmitted power fall within the aperture of the receiver. Beyond this the receiver finds itself in the hollow of the beam and there is a marked reduction in the coupling.

The coupling diagram may be imagined as the result of scanning the radiation pattern by a small aperture containing the sine-like curve of the receiver's complementary field. The scanning aperture diminishes in relation to the radiation pattern as range increases. At extreme range the receiver registers with virtually linear segments of the radiation pattern and its response depends on the slope of the segment. Thus at infinite range the coupling diagram is the ordinary polar diagram differentiated with respect to the polar angle.

Near the origin $G(u)$ is asymptotic to a straight line, which implies that the strength of the radiated electric field is directly proportional to the angle subtended at the transmitter by points near the axis. The same behaviour is to be expected of an exact solution. Thus the electric field in which the edge of the receiving aperture is situated is a function both of range and bearing. If the paraboloids are coaxial and the range is doubled, the bearing from the edge is halved. It follows that the electric field

incident upon the receiver diminishes inversely with the square of the separation. This effect prohibits the useful employment of the coaxial arrangement beyond the critical range.

(3) EXPERIMENTS WITH TWO DEEP PARABOLOIDS OF REVOLUTION

(3.1) Apparatus

The paraboloids are of $\frac{1}{16}$ in spun aluminium with a strong rim $\frac{1}{2}$ in thick. The profile, which is accurate to at least $\frac{1}{16}$ in, is proportioned as in Fig. 1 with dimension a equal to 4 in. Thus the radius of the aperture and the depth from apex to aperture plane are both 16 in. The appearance is that of outsize car headlamps rather than aerial reflectors, which are invariably shallow. If the surfaces had been terminated in the focal plane, 50% of the primary radiation would have been intercepted. As it is, 85% is intercepted, but to increase this figure still further would require a disproportionate increase in the amount of metal.

The dipole and its feeder are made from $\frac{1}{8}$ in diameter rigid brass tube, which carries the insulation and inner conductor of a coaxial line through a hole at the paraboloid apex. The inner conductor terminates on a 0·653 cm length of the tube, making half of the dipole. This is rather less than $\frac{1}{4}\lambda$ at 3·3 cm, to allow for the thickness of the dipole. The other half of the dipole is the outside of the feeder tube backed by a $\frac{1}{4}\lambda$ sleeve. An arrangement of platforms held by three taut wires from the rim positions the feeder tube behind the apex.

At the transmitter is a square-wave amplitude-modulated klystron. At the receiver detector, audio-frequency amplifiers and a valve voltmeter provide the sensitivity.

(3.2) Polar Diagrams

In Fig. 6 is shown the ordinary experimental radiation pattern of one of the paraboloids. This is taken using a horn as receiving

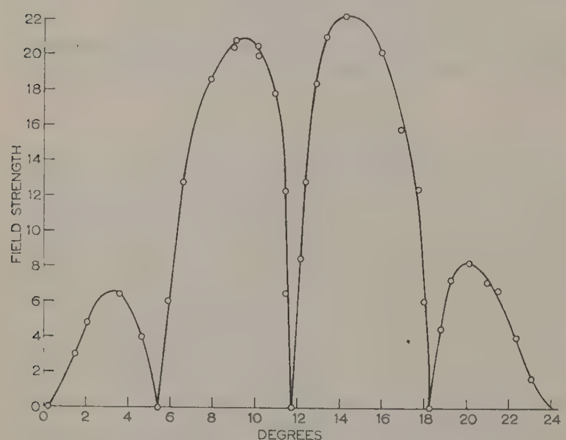


Fig. 6.—An experimental radiation pattern.

aerial in preference to the other paraboloid with misalignment. A third pair of lobes of significant amplitude is predicted by the approximate theory, but these are not found to exist. Otherwise the experimental diagram is substantially as expected.

Starting from axial alignment at the critical range, and turning one of the two paraboloids, the coupling diagram of Fig. 7 is plotted. The apparent half-power beam-width is 3° 18′ and no side-lobes beyond the first pair are observed. Similar results at

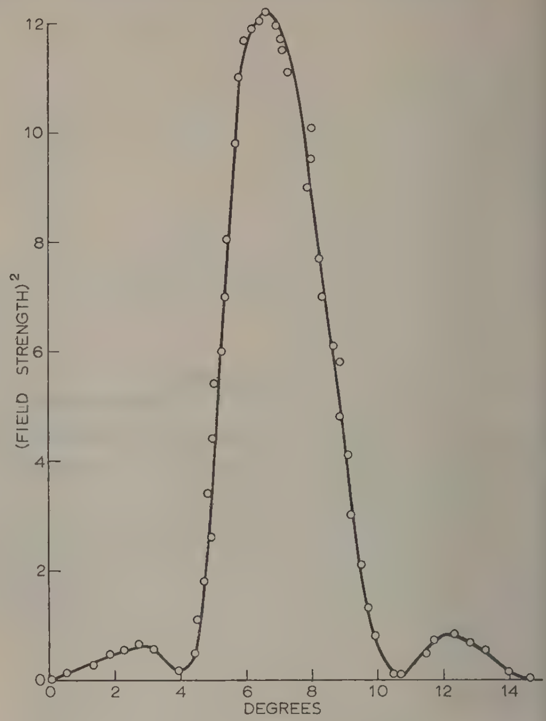


Fig. 7.—An experimental coupling diagram at 5 m range.

other ranges are 2° 30′ at 750 cm and 1° 55′ at 250 cm. The figures compare with an expected width of 2° 38′ from the co volution recorded in Fig. 5 and are in marked contrast wi the actual width of the hollow beam, which exceeds 10°.

(3.3) Variation of Coupling with Range

An estimate of the efficiency of transmission may be made as to exclude losses in the feeder cables and in the resistance the dipoles:

(*a*) The complete aerials are aligned physically for maximu coupling. Each aerial takes a turn as transmitter in order th matching to the feeder cables may be performed conveniently. matched detector is then fitted to the receiver and the amplifi output from it is recorded.

(*b*) The dipoles are separated from the paraboloids by sliding o their feeder cables. They are then employed as transmitter an receiver directly with normal orientation, and the range is adjus until the detector delivers the former output, after matching h again been checked.

(*c*) The original arrangement is restored to assess the repro ducibility of results.

A typical result is that the dipoles alone, separated by 8 mm give the same transfer of power as they achieve at 2 570 mm when placed at the foci of the paraboloids. This correspond to an overall power gain of more than 10^5 with respect to pair of unaided dipoles at that range. The coupling betwee the dipoles may be calculated in terms of the inverse-squar law and the absolute gain G of a dipole:

$$\frac{P_R}{P_T} = \left(\frac{G\lambda}{4\pi r}\right)^2 \qquad . \quad . \quad . \quad . \quad . \quad . \quad ($$

Thus, without assuming an inverse-square relation for th paraboloids themselves, their efficiency of transmission may b estimated:

paration between aperture planes	1	2	5	10	20	30 metres
iciency of transmission	70	24	14	3·3	0·6	0·1%

The worst experimental error of ±4% is in the first result, iere the coupling is high and the initial trim is difficult to produce exactly. If the whole of the power carried by the ain lobes were transferred, efficiencies of over 83% would be pected from the approximate theory. A curve of relative wer received against separation is shown in Fig. 8. Beyond e critical region of 5 m the power decreases more rapidly than would for any normally polarized system, which would obey

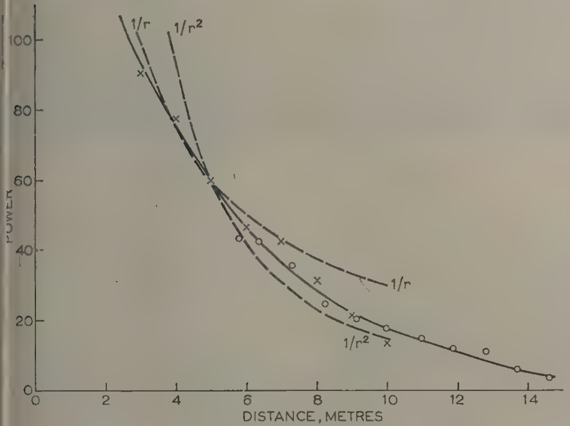


Fig. 8.—Variation of coupling with separation.

e inverse-square law. With axial excitation, field strengths are pected ultimately to diminish as $1/r^2$ and powers therefore as r^4. The observed diminution is not quite as rapid as this, it earth reflections are suspected of raising the received power the larger separations.

(4) DISCUSSION

The deep paraboloid with an axial source is an object of auty and symmetry. Mathematical solutions of the field uations within it are a whole order simpler than solutions for transverse source. Even so, an exact expression for the erture distribution requires Bessel functions, and the radiation ttern promises to be at least as complex. However, we can reasonably certain that both aperture distribution and diation pattern are odd functions which do not deviate arkedly from those associated with one cycle of a sine wave.

Two aligned paraboloids provide an example of reception iere the incident field is not a uniform plane wave. The riation of coupling is estimated by a process of convolution tween the aperture distribution of one aerial and the radiation ttern of the other projected on to the remote aperture. A upling diagram, which is notably different from and sharper an the ordinary polar diagram, is predicted and verified perimentally.

The final directivity is, however, no greater than that of iformly illuminated apertures of the same size. Indeed, the actical prospects for axial excitation are severely limited, cause the field near the axis diminishes so rapidly (as $1/r^2$ for fixed displacement from the axis). Only inside a critical nge, which allows the main lobes to be intercepted, is strong upling experienced. Here, there may possibly be latent plications to efficient power transfer and to short communica-on links with relative freedom from plane-polarized interference.

(5) ACKNOWLEDGMENTS

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(7) APPENDIX

Solution of Maxwell's Equations in Paraboloidal Co-ordinates[7]

Co-ordinates u, v, ϕ are taken, such that curves of u and v constant are confocal paraboloids with focus at the origin and curves of ϕ constant are intersecting planes on the axis of revolution. They are related to cylindrical co-ordinates r, ϕ, z thus:

$$z + jr = l(u + jv)^2 \quad . \quad . \quad . \quad . \quad . \quad (9)$$

From the symmetry of an axial excitation system and from experience with similar cylindrical cases, we know that the only component of H is H_ϕ, which must satisfy the equation

$$\text{curl curl } H_\phi = \omega^2\epsilon_0\mu_0 H_\phi \quad . \quad . \quad . \quad . \quad (10)$$

Let us write the components of curl H in terms of a vector G:

$$G_u = \text{curl}_u H = \frac{1}{2lv\sqrt{(u^2+v^2)}}\frac{\partial}{\partial v}(vH_\phi) - \frac{1}{2luv}\frac{\partial Hv}{\partial \phi} \quad . \quad (11)$$

$$G_v = \text{curl}_v H = \frac{1}{2lv}\frac{\partial H_u}{\partial \phi} - \frac{1}{2lu\sqrt{(u^2+v^2)}}\frac{\partial}{\partial u}(uH_\phi) \quad . \quad (12)$$

$$G_\phi = \text{curl}_\phi H = \frac{1}{2l(u^2+v^2)}$$
$$\left[\frac{\partial}{\partial u}\sqrt{(u^2+v^2)}H_v - \frac{\partial}{\partial v}\sqrt{(u^2+v^2)}H_u\right] \quad . \quad (13)$$

The expansion of [curl $G]_\phi$ has the form of eqn. (13) and requires substitution from eqns. (11) and (12):

$$\text{curl } G_\phi \quad \frac{-1}{4l^2(u^2+v^2)}\left[\frac{\partial}{\partial u}\left(\frac{H_\phi}{u} + \frac{\partial H_\phi}{\partial u}\right) + \frac{\partial}{\partial v}\left(\frac{H_\phi}{v} + \frac{\partial H_\phi}{\partial v}\right)\right]$$
$$. \quad . \quad . \quad . \quad . \quad (14)$$

This is curl curl H and may be equated with the right-hand side of eqn. (9) and expanded thus:

$$\omega^2\epsilon_0\mu_0 H_\phi = -\frac{1}{4l^2(u^2+v^2)}$$

$$\left(\frac{-H_\phi}{u^2} + \frac{1}{u}\frac{\partial H_\phi}{\partial u} + \frac{\partial^2 H_\phi}{\partial u^2} - \frac{H_\phi}{v^2} + \frac{1}{v}\frac{\partial H_\phi}{\partial v} + \frac{\partial^2 H_\phi}{\partial v^2}\right) \quad . \quad (15)$$

To separate the variables we now postulate a trial solution in the form $H_\phi = UV$, where U is a function only of u, and V is a function only of v.

Identical separate equations are found for U and V:

$$\frac{\partial^2 V}{\partial v^2} + \frac{1}{v}\frac{\partial V}{\partial v} + \left(4\omega^2\epsilon_0\mu_0 l^2 v^2 - \frac{1}{v^2}\right)V = 0 \quad . \quad (16)$$

This equation has solutions in the form of power series, but it is so reminiscent of Bessel's equation that the following substitutions are tried.

Let $\quad y = V$ . . . .

$x = \frac{1}{2}kv^2$ . . . . . .

where $\quad k = 2\omega\sqrt{(\epsilon_0\mu_0)}l$ . . . . .

Then $$\frac{dy}{dv} = \frac{dy}{dx}\frac{dx}{dv} = kv\frac{dy}{dx} \quad . \quad . \quad . \quad . \quad .$$

and $$\frac{d^2y}{dv^2} = k^2v^2\frac{d^2y}{dx^2} + k\frac{dy}{dx} \quad . \quad . \quad . \quad .$$

Eqn. (16) is multiplied throughout by v^2 and results (20) (21) are applied:

$$v^4k^2\frac{d^2y}{dx^2} + v^2k\frac{dy}{dx} + v^2k\frac{dy}{dx} + v^4k^2y - y = 0 \quad . \quad .$$

Making use of eqn. (18) and dividing throughout by 4, arrive at a standard form of Bessel's equation of order one-h

$$\frac{d^2y}{dx^2} + \frac{1}{x}\frac{dy}{dx} + \left(1 - \frac{1}{2^2x^2}\right)y = 0 \quad . \quad . \quad .$$

A PHYSICAL CLASSIFICATION OF ELECTROMAGNETIC WAVES

By Prof. H. E. M. BARLOW, Ph.D., B.Sc.(Eng.), Member.

(*Communication received* 24*th August*, 1960)

In homogeneous media it is possible to recognize in the broadest sense three and only three distinctive types of electromagnetic wave. These are (i) the TEM wave in which, as the name implies, there are no longitudinal components of either the electric or the magnetic fields, which remain constant in phase over the wavefront, (ii) the surface wave in which at least one of the field components over the wavefront is evanescent, suffering continuous decay in amplitude without change of phase, and (iii) the so-called waveguide modes characterized by a transverse standing wave over the wavefront. It will be apparent that by 'wavefront' is meant in all cases an equi-phase surface.

The first of these types is usually regarded as only associated in pure form with free-space propagation, but if we accept the description TEM as applying to any given point over the wavefront, instead of over the wavefront as a whole, guided waves of this kind supported by twin-conductor transmission lines are also included in the group. It must be recognized, however, that with any such guided-wave system there are two boundary conditions to satisfy, and consequently a waveguide mode is also possible when the spacing of the conductors is greater than half a wavelength. Moreover, if one conductor has a highly reactive surface, a close approximation to a surface wave may be formed in proximity to that conductor, the corresponding field decaying in amplitude so rapidly with distance from the boundary surface concerned that the second conductor has no very significant effect.

The fact remains, however, that a guide consisting of two independent parallel conductors is, in general, capable supporting a hybrid TEM/surface wave, and when there sufficient spacing between the conductors a superimpo 'waveguide' mode.

The surface wave can be conveniently regarded as takin place in our classification which is intermediate to the TEM a the waveguide modes. Thus the TEM wave has no wa distribution of any kind over the wavefront, the surface wa has an evanescent distribution and the waveguide mode has standing wave. As a consequence of this classification it is on the waveguide mode that is subject to a cut-off. Strictly speakin a surface wave cannot exist as a pure mode if there is more th one finite boundary condition to satisfy. As a corollary, t waveguide mode requires at least two boundaries, and the may be represented by interfaces between the propagati medium and either two different parts of the same guide or tw distinct structures which together form the guide.

Most guided-wave systems are designed to give predomina support to one of the three distinctive types of wave giv separate classification here, but, in fact, a hybrid wave is genera produced incorporating a mixture of all three. If the co ditions are suitable for the support of a waveguide mo this tends to be more strongly excited because of its resona nature.

In analysing the behaviour of a given supporting structure a waveguide, it is therefore of special interest to resolve t field equations, if possible, into three parts, representing t three types of wave separately recognizable by the distincti characteristics suggested.

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Nov. 1960

TELEPHONE ECHO TESTS

By D. L. RICHARDS, B.Sc.(Eng.), and G. A. BUCK, Associate Members.

(*The paper was first received* 22*nd February, and in revised form* 29*th June,* 1960.)

SUMMARY

Electrical reflection on telephone circuits having appreciable times [of] propagation can cause delayed sidetone effects subjectively similar [to] echo. The variations in subjects' tolerance of this echo is examined [ex]perimentally as a function of echo attenuation and delay. The dis[pe]rsion of tolerance between subjects is also considered, and this [yi]elds information in a form suitable for planning telephone networks [so] that echo will not cause excessive inconvenience.

(1) INTRODUCTION

Reflections in telephone circuits, principally due to necessary [c]ompromise balancing at 4-wire/2-wire hybrid connections, can [re]turn speech signals to the sending end. If the propagation [t]ime is appreciable the reflected signals, if not sufficiently [a]ttenuated, may give rise to a delayed sidetone effect perceptible [to] the talker and termed 'echo'. Multiple reflections similarly [m]ay give rise to spurious delayed signals which the listener may [h]ear at the receiving end. In practice if talker echo, as the [fo]rmer is called, is reduced to acceptable proportions, listener [e]cho, being generally subject to greater attenuation, is unlikely [to] cause trouble. The present investigation is concerned only [w]ith talker echo.

When echo is present on a telephone connection it will have [so]me adverse effect upon the opinion of the customer using it. [N]ot all subjects will be inconvenienced to the same extent, some [b]eing more susceptible than others. Even a call between two [gi]ven stations may encounter, on different occasions, a different [de]gree of echo owing to unavoidable variations in the plant. [In] general, a telephone network must be planned so that echo [w]ill not be present to a deleterious extent except in a very small [p]roportion of calls. To achieve a design which will ensure [th]is desirable state of affairs, information must be available on [th]e dispersion of tolerance to echo among customers as well [as] knowledge of the extent of variation of physical charac[te]ristics of the telephone network relevant to the production of [ec]ho.

Of the prior information[2,3] available, the C.C.I.F. curve [(s]ee Fig. 1) relates permissible echo-path attenuation to delay [fo]r the connexion between the terminal exchanges involved in a [tr]unk telephone call. No information is available concerning [th]e degree of discomfort or inconvenience involved nor is the [di]spersion of the subjects' tolerance revealed. Such information [in] the form of a 'limit' could be applied when a particular indi[vi]dual physical connection was being considered; it is not [a]daptable for the broader purposes of comparing the relative [m]erits of different complete networks (each consisting of an [ag]gregate of connections).

The Bell Telephone Laboratories curve (see Fig. 1) is based [o]n the level of talker echo to be 'commercially tolerable', i.e. [di]scernible but not objectionable. The standard error is quoted [...] dB), but the number of subjects and other details are not, so [th]at it is not possible to deduce the 'between subjects' standard [de]viation.

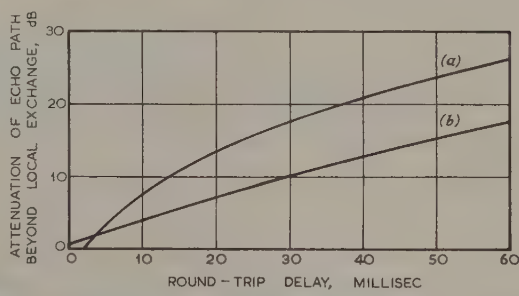

Fig. 1.—Relationship between minimum permissible echo-path attenuation and round-trip delay.

(*a*) Bell Telephone Laboratories.
(*b*) C.C.I.F. (now C.C.I.T.T.) proposals.

(2) QUANTITIES MEASURED EXPERIMENTALLY

In measuring the effects of talker echo in the laboratory, chief reliance was placed upon the opinions expressed by subjects after they had read a set passage or otherwise spoken into a telephone connected to apparatus simulating echo but not connected to a distant party. After each reading each subject was asked to say whether or not echo effects were perceptible, and if so, whether or not they would be thought objectionable if encountered on an ordinary telephone call.

An opinion lying within one or other of the three categories 'not perceptible', 'perceptible but not objectionable' and 'objectionable' was thus obtained for each combination of subject and circuit condition tested. By varying one or more physical stimuli (usually echo-path sensitivity*) between readings and observing how the opinions were affected, estimates in terms of circuit parameters were obtained of the inter-category boundaries corresponding to 'just perceptible' and 'just objectionable' echo.

The speech voltage produced in the telephone circuit by each talker was measured, and also, in some experiments, the time taken to read the set passage. The way in which these quantities were affected by changes in stimuli could thus be examined.

One experiment was of a conversational type,[5] and in this subjects conversed with one another in pairs by means of a telephone circuit in which talker echo could be simulated.

(3) MISCELLANEOUS EXPERIMENTAL FEATURES

In all the experiments undertaken, talker echo phenomena were simulated by means of a magnetic-drum audio-delay device capable of producing delays having any value between 8 and 64 millisec.

Most tests incorporated a laboratory telephone circuit, specially built for subjective experimental work, which terminates on handsets using electromagnetic earphones and microphones. Although similar in appearance and feel to an

[W]ritten contributions on papers published without being read at meetings are [inv]ited for consideration with a view to publication.
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* In the testing of speech links interest is focused on complete paths from the talker's mouth to his correspondent's ear or, in the case of sidetone or echo, to his own ear. It is convenient to consider the performance of such a path or any part of it in terms of sensitivity rather than attenuation because it consists generally, not only of an electrical portion, but also of acoustical portions and transducers.

ordinary telephone so far as the subjects are concerned, its response is free from non-linear distortion and in the range 300–3 400 c/s is independent of frequency. It is basically a 4-wire system, and independent adjustments to experimental stimuli, notably the sensitivities of the sidetone and echo paths, were accordingly simplified.

Other tests were conducted using a commercial telephone circuit typical of the British Post Office network; in those cases telephone sets Type 13-2P-27 were used with either zero or average subscribers' lines.

In some of the talking tests a key was provided by means of which each subject could at will connect or disconnect the echo simulating path as an aid in deciding whether or not echo effects were present. In comparison with tests in which no such key was used, however, neither the level of the threshold of perception of echo nor the standard error of its measurement was significantly reduced.

All the talking tests were performed in the absence of room noise, this being the most adverse ambient condition from the point of view of echo. The conversational test attempted to simulate average conditions, and a room noise of typical spectrum was used at a level of −50 dB measured with an American sound-level meter. The effects of room noise were not thought likely to be very large and were not, therefore specifically studied.

(4) PHYSICAL STIMULI STUDIED AND THEIR EFFECTS ON SUBJECTIVE OPINIONS

In all the experiments conducted, the primary controlled variable was the air-to-air sensitivity of the overall echo path. In the practical situation this path would comprise the sending direction of the telephone circuit, the line, the remote unbalance and so back to the talker in the receive direction. The sensitivity was adjusted in random sequence to each of a range of prearranged values, observations being made at each setting. Other sets of observations were then made using different settings of other circuit parameters.

Fig. 2 shows the results of a series of sets of observations of subjects' perception of and objection to echo conducted at a number of values of echo delay. It can be seen that, as the

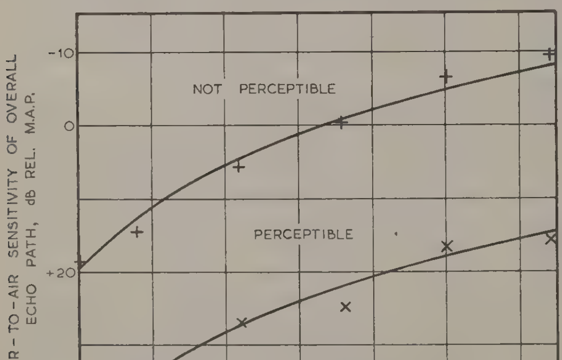

Fig. 2.—Effect of round-trip delay on median inter-category boundaries of talker echo, using laboratory telephone.

Data for zero delay were obtained by varying sidetone sensitivity. At other delays, sidetone sensitivity was +13 dB rel. m.a.p. (i.e. relative to the setting giving a loudness equivalent to that of a metre air path.[5])

sensitivity of the echo path is increased, echo becomes fi perceptible to the average individual and then objectionat As the delay is increased the sensitivities at which percepti and objection occur are reduced. This is in qualitative agr ment with the previously published information already refer to.

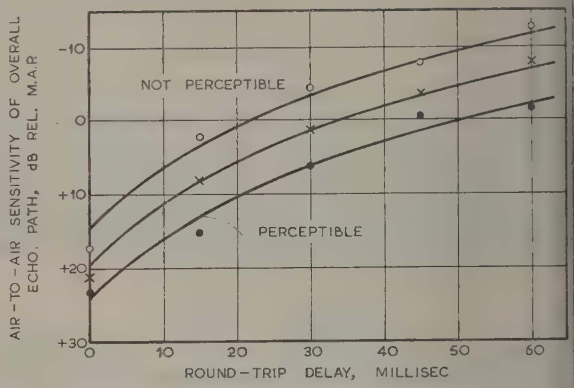

Fig. 3.—Effect of round-trip delay and sidetone sensitivity on the mean thresholds of perception of talker echo.

○ Sidetone sensitivity, +10 dB rel. m.a.p.
× Sidetone sensitivity, +20 dB rel. m.a.p.
● Sidetone sensitivity, +30 dB rel. m.a.p.

Fig. 3 shows how the thresholds of perception of echo articulation crew subjects* change as a result of changes both echo delay and sidetone sensitivity when the comparativ high-quality laboratory telephone is used. An increase in si tone sensitivity of 10 dB raises the threshold of perception terms of echo-path sensitivity by about 5 dB. In other word echo tends to be masked in the presence of appreciable sideto The relationship, although linear in the range shown and in pendent of round-trip delay, is not valid for sidetone sensitivit outside this range. This was verified experimentally for t case of 25 millisec round-trip echo delay. Other work has sho that quieter sidetone becomes inaudible to an increasing numb of people and thus must have a diminishing masking effect echo; louder sidetone tends to produce objection to the sideto itself on the ground of excessive loudness irrespective of ec effects, and consequently the opinion category of echo 'perce tible but not objectionable' becomes inappropriate.

The overall send-plus-receive sensitivity of a commerc telephone as used in the tests decreases uniformly with lo line length. Thus an increase in line length is equivalent to reduction in air-to-air sensitivity of the echo path. It w found, however, that a change in subscribers' line from zero one equivalent to a mile of 6½ lb/mile cable had no observa effect on the threshold of perception of echo when this w expressed in terms of attenuation of the echo path from t local exchange to the point of reflection and back to the lo exchange. The change in sensitivity of the sidetone path normally measured between these two line conditions is inad quate to account for this result unless, owing perhaps to chang in distortion, the changes in the echo-masking capabilities sidetone can be greater than its apparent loudness would le one to expect. At greater lengths of local line than the one m actually used in the tests, sidetone sensitivity changes mu less rapidly although overall sensitivity continues to fall. It therefore reasonable to assume that telephones on these long

* One experiment explored a rather large number of combinations of stimuli a required the services of each subject for a considerable time. To avoid inconvenie to volunteer untrained subjects who would otherwise have taken part, articulat crew members were employed. Their perception of echo was slightly more ac than that of untrained subjects and their dispersion slightly less.

ıes will at least be generally no more susceptible to echo. In edicting the effects of echo in the public network it seems erefore permissible, and is convenient, to disregard the length subscriber's line and consider only the remainder of the ho path lying beyond the local exchange.

Another feature of the commercial telephone is that a given ange in round-trip delay produces a slightly greater change the level of the threshold of perception than is obtained with e laboratory telephone.

These facts suggest that caution should be exercised in tending the present results to other types of telephone set. The median threshold of objection to echo was not directly vestigated on a commercial telephone except at 60 millisec lay. At lower values of round-trip delay this would require echo-path gain which is somewhat difficult to simulate lidly. Although lower percentile thresholds of objection uld in part be explored without an echo-path gain, a rather rmidable amount of experimental effort would be required to tain reasonable accuracy. The median threshold was therere inferred from data on the threshold of perception using the ter-threshold interval obtained on the laboratory telephone.

(5) DISPERSION OF SUBJECTS

The opinion data collected in the talking tests showed that me untrained subjects were much more disturbed than others a given level of echo signals. The distribution of tolerances subjects about the median threshold of objection was found all cases to be substantially Gaussian, and estimates of the ındard deviation of this distribution were obtained by probit alysis.[1] One experiment using 10 untrained subjects gave a lue of 8 dB, and another, 13 dB. An experiment using 36 bjects who were expected to have some relevant technical owledge gave 9 dB. In the absence of considerable additional ›rk an average value of 10 dB seems reasonable and has been opted.

The possible effects of talker volume, or speech level, on echo enomena were not studied. Talker volume cannot very ıdily be controlled by engineering techniques and varies dely in the individual from time to time; between individuals variation is even greater. Any such effects are therefore ›uped with other variations between subjects in arriving at overall dispersion of echo tolerance.

(6) PLANNING PROPOSALS

The family of curves (*a*)–(*e*) in Fig. 4 was generated by applying standard deviation of the distribution of subjects to the dian threshold of objection to echo when using a commercial- ›e telephone. The median is given in terms of the round-trip echo-path attenuation as a function of round-trip delay. The minimum permissible echo-path attenuation (round trip from local exchange back to local exchange) to ensure that not more than a given percentage of subjects will find talker echo objectionable can now be found for any value of delay up to 60 millisec, and thus any desired grade of service from the point of view of echo can be secured.

The planning limits thus obtained are, however, subject to the reservation that they are based on opinions expressed by subjects whose attention was expressly directed to echo. Moreover, each subject was reading into a telephone and not conducting purposeful conversation. Some reassurance on these features is forthcoming from the conversational test discussed later, but in applying the limits it is difficult to avoid the assumption, albeit perhaps not unreasonable in view of the subjects used here, that the behaviour of the subjects is representative of that of the population of telephone users.

(7) THE CONVERSATIONAL OPINIONS EXPERIMENT

A conversational opinions experiment consists of getting subjects to converse in pairs over telephone links to be assessed. Pictorial puzzles are used to stimulate purposeful conversation, and at the end of each telephone call each subject is asked to give his opinion of it by reference to a prearranged scale. The scale used in the present instance consisted simply of five categories 'excellent', 'good', 'fair', 'poor' and 'bad'. In analysing the results of the experiment these opinions were scored numerically 4, 3, 2, 1 and 0 respectively. By averaging the opinions of all the subjects for each treatment, or setting of circuit parameters, a mean opinion score for each treatment is found.

The results obtained were of little help in directly formulating planning limits because the data defined no precise thresholds at which echo could be said to commence to have a significantly depressing effect on opinions. Nevertheless, although many of the differences are not statistically significant, a clear trend towards less favourable opinions is produced as echo becomes less attenuated or more delayed. Table 1, cols. 2–7, shows some of the results. Each value shown is subject to a 95% confidence interval of $\pm 0{\cdot}5$ units.

In a conversational experiment of the size used (13 pairs of subjects) it is not possible to obtain directly a reliable estimate of the percentage of 'unsatisfactory' calls, i.e. those declared 'poor' or 'bad'. It has been found, however, that a reasonably stable relationship exists between such percentages and the mean opinion score.[4] Applying this to the mean opinion scores in cols. 2–4 gives the percentages quoted in cols. 8–10. Cols. 11–13 show the percentages of subjects finding echo objectionable after reading a set passage into a telephone which produces echo; these were obtained from the proposed planning data

Table 1

Results of Certain Treatments in the Conversation Opinions Experiment

Echo-path attenuation	Mean opinion score (m.o.s.)						Percentage of subjects giving opinions 'poor' or 'bad' derived from m.o.s. for 10 dB junction loss			Percentage of subjects finding echo 'objectionable' (from Fig. 4)		
	Junction loss, 10 dB			Junction loss, 20 dB								
	Delay 10 ms	25 ms	40 ms	10 ms	25 ms	40 ms	10 ms	25 ms	40 ms	10 ms	25 ms	40 ms
dB												
(No echo)	3·35	3·35	3·35	2·71	2·71	2·71	0	0	0	0	0	0
25	3·70	3·00	2·47	2·70	2·96	2·51	0	0·8	8·0	0	1	4
20	2·93	2·62	2·55	2·97	2·31	2·13	1·1	3·7	6·1	0	2	10
15	2·59	2·62	2·24	2·36	2·27	2·09	5·2	3·7	15·3	1	6	21
10	2·66	2·23	1·59	1·93	1·46	1·20	4·0	15·7	47·7	2	15	38

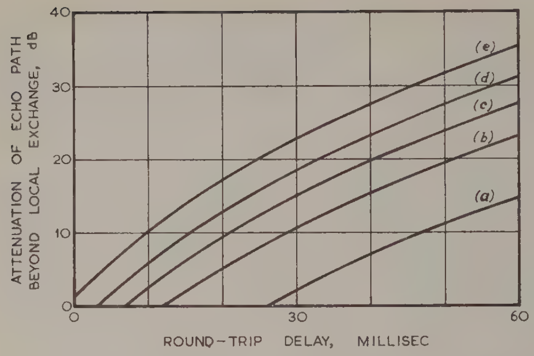

Fig. 4.—Proposed relationships between echo-path attenuation and round-trip delay based on present investigations.

Number of subjects objecting to echo:
(a) 50%
(b) 20%
(c) 10%
(d) 5%
(e) 2%

Measurements of both echo-path attenuation and delay relate to that part of the echo path existing from the local exchange to the point of return and back to the local exchange.

given in Fig. 4. Rough agreement is seen, but this would apply only in the case in which a negligible percentage of subjects regard the circuit used for conversation as 'unsatisfactory' owing to degradations other than echo.

(8) SPEECH VOLTAGE AND READING TIME

The ways in which the speech voltage produced in the telephone circuit and the time taken to read a set passage are influenced by echo-path sensitivity and delay are shown in Fig. 5.

At delays smaller than 30 millisec, speech voltage is noticeably depressed at higher values of echo-path sensitivity just as it would be by excessive sidetone. No effect is discernible at longer delays. The reading time, on the contrary, is unaffected by any reasonable level of sidetone or echo when the latter is delayed by less than 30 millisec. But at greater delays an increase in either echo delay or sensitivity results in an increase in reading time.

Although no corresponding discontinuities appear in the opinions data, these results suggest that echo signals are in some way 'sidetone-like' at delays of less than about 30 millisec and 'echo-like' at greater values. Subjects in other work[6] have been observed to appreciate delayed signals as discrete echoes only at delays exceeding about 30 millisec: the experimental situation in that case, however, corresponded rather to listener than to talker echo conditions.

(9) CONCLUSIONS

The planning proposals obtained are not based on as much experimental data as could perhaps be desired, nor do they extend to echo delays greatly exceeding 60 millisec. They do, however, take account of the dispersion of echo tolerance between people and thus permit limits of performance of a network from the point of view of echo to be specified in order to obtain a desired grade of service. Alternatively, for a network of known or assumed echo performance, the probability of dissatisfaction arising from echo can be estimated.

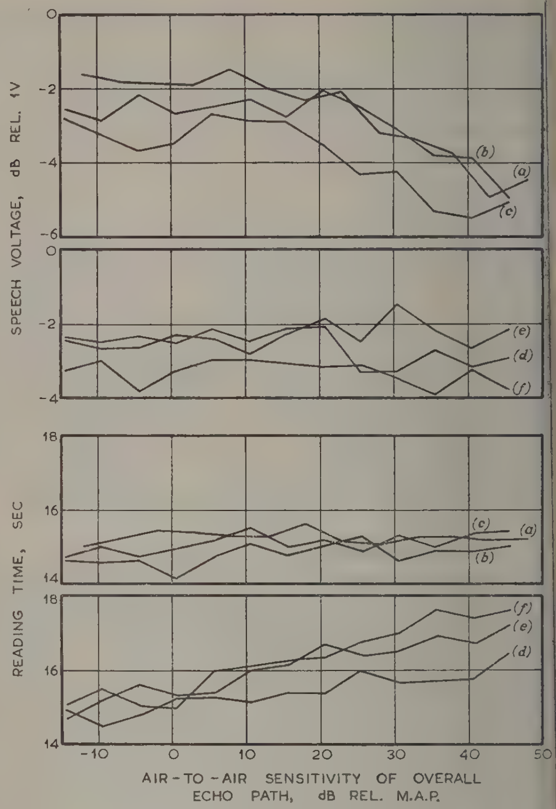

Fig. 5.—Speech voltage and reading time as a function of echo-p sensitivity and delay.

(a) Delay, zero. Sidetone path sensitivity varied.
(b) Delay, 8 millisec.
(c) Delay, 22 millisec.
(d) Delay, 36 millisec. } Sidetone path sensitivity, +13 dB rel. m.
(e) Delay, 50 millisec.
(f) Delay, 64 millisec.

(10) ACKNOWLEDGMENTS

Acknowledgment is made to the Engineer-in-Chief of Post Office for permission to make use of information contai in the paper.

The authors' thanks are extended to their colleagues who t part in the subjective experiments, and to Mr. C. J. Skin who conducted them.

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INDUSTRIAL, BIOLOGICAL AND MEDICAL ASPECTS OF MICROWAVE RADIATION

By A. F. HARVEY, D.Phil., B.Sc.(Eng.), Member.

(*The paper was first received* 12*th October*, 1959, *and in revised form* 11*th April*, 1960.)

SUMMARY

The paper reviews the industrial, biological and medical aspects of icrowave radiation. The special methods of study of the properties ' organic and biological materials are first discussed. The industrial oplications of microwave heating processes are described and the fect of microwave radiation on biological tissues and living animals examined. The operational hazards attaching to personnel in the ighbourhood of high-power equipment are pointed out and suggesons offered as to how these can be minimized. The paper concludes ith a bibliography.

LIST OF PRINCIPAL SYMBOLS

Rationalized M.K.S. units are used unless otherwise indicated.

d = Diameter, m.
E = R.F. peak electric field strength, volts/m.
f = Frequency, c/s.
g = Spectroscopic splitting factor or g-factor = $2 \cdot 002\,29$ for a free electron.
P = Power, watts.
P_s = Power dissipated in sample, watts.
R = Range, m.
S = Poynting vector of power flux density = $E^2/754$ watts/m^2.
t = Time, sec.
V_s = Volume of sample, m^3.
α = Attenuation coefficient, nepers/m (=$8 \cdot 686$ dB/m).
δ = Dielectric loss angle = arc tan (ϵ''/ϵ').
ϵ = Relative permittivity = $\epsilon' - j\epsilon''$.
λ = Free-space wavelength, m.
σ = Conductivity, mhos/m.
$\cos \phi$ = Power factor $\simeq \tan \delta$ if $\delta \ll 1$.

(1) INTRODUCTION

Microwaves are defined to be that portion of the electromagetic spectrum with frequencies between 1 Gc/s (wavelength)cm) and 100 Gc/s (wavelength 1 mm). Originally developed n a large scale for radar systems, microwave techniques have nce been applied to other fields, including radiocommunication, ielectric spectroscopy, radio astronomy and electro-nuclear cceleration. Electromagnetic energy penetrates bodies to an xtent governed by the properties of the material and the freuency and other characteristics of the incident waves. This ontrolled penetration forms the basis of the industrial, biogical and medical applications of microwave radiation.

The effects observed under exposure to microwaves are, to a rge extent, thermal: in this case it is the dispersive properties of ermittivity and conductivity which play essential roles. On the ther hand, effects specifically related to the frequency of radiaon have been observed: such cases involve the molecular structure, which may be studied by electron-resonance techniques. Dielectric heating forms the basis of industrial manufacturing processes and the defrosting or cooking of foods. The effect of intense microwave radiation on living animals depends upon the circulation of blood, and thus investigations must be made on tissues with and without blood vessels. These short- and long-term effects concern humans who must work in positions of intense microwave fields. Methods of reducing such hazards involve instrumentation for measurement of the radiation intensity and adoption of efficient methods of protection.

Some of these industrial, biological and medical aspects of microwave radiation are relatively new but, although development is proceeding rapidly, accurate measurements have been inadequate with the result that reliable conclusions cannot be drawn. The paper thus takes the form of a factual outline with occasional detail to illustrate the problems involved. The bibliography will enable the reader to supplement and amplify the subject matter to give a balanced and complete coverage of these important and topical fields.

(2) PROPERTIES OF BIOLOGICAL MATERIALS

(2.1) Dielectric Behaviour

The dielectric properties of biological material may be examined[77, 83] by methods closely resembling those employed with ordinary materials: in many cases the dispersions obtained follow conventional laws. For example, the dielectric constant of blood shows the characteristics expected of its high water content.[19, 73, 82] The results, together with those for tissues such as skin, muscle, bone and fatty parts, measured in the range $3 \cdot 0$–$23 \cdot 6$ Gc/s[30, 31] and $1 \cdot 78$–$4 \cdot 63$ Gc/s,[17] can be fitted to a Debye relation.[18] Typical data, selected from measurements[45, 84, 94] on various human tissues in the frequency range $0 \cdot 2$–$10 \cdot 0$ Gc/s, are given in Table 1. The differences in the properties are to some extent due to the presence of protein molecules.[44] The temperature coefficient of the resistivity varies with frequency, is always negative and is comparable with that of saline solution.[20]

Table 1

PROPERTIES OF VARIOUS TISSUES AT 27° C

Tissue	Relative permittivity at		Conductivity at	
	0·4 Gc/s	1 Gc/s	0·4 Gc/s	1 Gc/s
			millimhos/cm	millimhos/cm
Muscle ..	54–56	54 57	9·5–10·0	12·0–12·7
Liver ..	46–53	50–51	6·7–8·3	9·1–9·6
Lung ..	36	—	6·1	—
Blood ..	64	63–67	11·0	12·8–14·2
0·9% NaCl	74	77	17·3	18·5
Fat	4–7	—	0·7–0·8	—

The temperature coefficient of the relative permittivity is posi-

This is an 'integrating' paper. Members are invited to submit papers in this tegory, giving the full perspective of the developments leading to the present actice in a particular part of one of the branches of electrical science.
Written contributions on papers published without being read at meetings are vited for consideration with a view to publication.
Dr. Harvey is at the Royal Radar Establishment.

Table 2

PERMITTIVITY AND CONDUCTIVITY OF TISSUES AT 37° C

Frequency	Muscle		Wet fat		Dry fat		Skin	
	ϵ	σ	ϵ	σ	ϵ	σ	ϵ	σ
Gc/s		millimhos/cm		millimhos/cm		millimhos/cm		millimhos/cm
0·15	66	10	7·6	0·66	3·8	0·33	63	8
0·4	58	10	6·8	0·78	3·4	0·39	47	9
0·9	54	11	6·1	0·91	3·1	0·45	44	10
3·0	54	22	4·4	1·18	2·2	0·59	41	21
10·0	45	125	3·3	2·63	1·7	1·31	34	82

tive at low frequencies, and becomes zero and finally negative as the frequency increases. In view of their medical importance the properties of muscle, fat and skin have been studied over a wide range of frequency:[17, 30, 88] typical results for the relative permittivity and conductivity are given in Table 2. The 'wet' fat, with somewhat higher water content, represents horse fat, while the 'dry' fat has been found more characteristic of pork: the values for human fat are somewhere in between. The permittivity of wet fat does not change much with temperature but the conductivity, as compared with the value at 37° C, is doubled at 50° C and halved at 20° C.

(2.2) Electron-Resonance Techniques

Electron-resonance methods[57] are proving useful in biological and medical studies since they facilitate examination of free radical reactions which are often associated with metabolic and biochemical processes.[5, 6, 7, 98] The presence of water tends to cause a large non-resonant absorption, and thus the specimens are usually enclosed in tubes of, say, 1 mm diameter. Alternatively, the samples can be freeze-dried, care being taken that no extra bonds are broken and radicals formed. Commoner, Townsend and Pake[14] have, for example, studied the free-radical concentration in living tissue and typical results are given in Table 3. The free radicals were shown to be associated

Table 3

FREE-RADICAL CONCENTRATION IN TISSUES

Material	Concentration (dry weight)
	mole/g × 10^{-8}
Nicotiana tabacum, leaf ..	65
Coleus, leaf	180
Barley, leaf	25
Digitalis, germinating seeds	10
Carrot, root	8
Rabbit, muscle	20
Frog, eggs	200
Drosophilia, entire ..	4

with the protein components and the concentration is seen to be higher in the metabolically active tissues. In the case of leaves the concentration could be increased on exposure to light.

Free radicals have been postulated[67] as taking part in carcinogenic (cancer-forming) processes and it has been suggested that the activity of certain large-ring structures may be related to their ability to form negative-ion free radicals with mild reducing agents. In contrast, non-carcinogenic hydrocarbons, such as naphthalene, require strong reducing agents before the formation of such radicals. Thus, carcinogenic activity might be measured by comparing the free-radical concentration in healthy cancerous tissue. For example, cigarette smoke, when formed, contains a relatively high concentration of both ac and stabilized free radicals,[62] either or both of which might as carcinogenic agents.

Free-radical processes are likely to occur in enzyme reactio the details of which can then be studied by electron-resona techniques. Closely related to biologically important molec are the phthalocyanines: these are large planar molec which can undergo a transient intermediate oxidation sta Phthalocyanines studied experimentally[37] show that this transi state gives a narrow resonance line with a g-factor close to t of a free-electron spin. This indicates that the oxidation p cess involves mobile electrons in the conjugated ring syst and, in some cases, the growth and decay of the intermedi stage can be followed on the oscillograph display of spectrometer.

The biologically important oxidation of ferrihaemoglobin its metastable state has been examined in detail[39, 40] by electr resonance methods. In fact, such study of the different deri tives of haemoglobin not only allows the actual orbitals invol in the binding to be determined but also gives detailed structu information of the processes involved. A strong narrow with a g-factor of 2·003 is obtained when methaemoglobin metmyaglobin is oxidized by, for example, hydrogen perox The fact that the formation of an actual peroxide compou involves a change in the binding of the iron atom is demonstra conclusively by observing[2, 56] the metmyoglobin resonance $g = 6\cdot0$ as oxidation occurs. The fact that the resona decreases[41] during the oxidation process indicates a change fr ionic to covalent binding.

Kinetic studies have shown the role played by free radi during photosynthesis. In one series of experiments aque suspensions of chloroplasts were examined[15] in a resonant ca into which light from a tungsten-filament lamp could be admitt this illumination increased by sixfold the radical concentrati Results at various temperatures[95] show a longer decay time —140° C, and correlation of luminescence and electron-resona measurements[97] indicate that the photosynthesis reaction is to the mechanism of electron trapping and hole product familiar in semiconductors.

Electron resonance affords a direct and sensitive method studying the breakdown processes which occur in living tis as a result of X- or γ-ray irradiation. Pulse techniques ena dynamic conditions to be studied. Results have shown, example, similar spectra for cystine, hair, nail and feathe These techniques can also be applied[55, 93] to the study of prote and such trace metals as manganese and copper. Experiment on mice have shown that paramagnetic tracers such as salts iron, nickel or chromium enable blood flow to be measur The technique involves determination of the spin-lattice rela

on time after application of a saturating pulse. The relative mplitudes of the decaying signal, measured when blood is owing and not flowing, gives the velocity of flow knowing the istance from the injection point. The method is safe and is uitable for use with humans.

(3) INDUSTRIAL HEATING PROCESS

(3.1) Methods Employed

Many methods have been developed for utilizing the dielectric oss, shown by most materials at ultra-high and microwave freuencies, for heating purposes.[42, 59, 64, 80, 81] The characteristic dvantage of dielectric heating at any frequency is its ability to roduce rapid uniform heating throughout the bulk of a material. The total power dissipated in the volume of the dielectric is iven by

$$P_s = 2\cdot78 \times 10^{-11} f E^2 V_s \epsilon'' \quad . \quad . \quad . \quad . \quad (1)$$

nd thus to increase the rate of heating of a given material, the pplied field strength or the frequency must be increased. The ormer is easy to control but can be increased only to the limit mposed by arcing between the electrodes and the work. The rcing limit varies considerably with different kinds of materials: n a substance of a discrete nature the concentration of field at he contact points leads to local burning at relatively low powers.

Eqn. (1) shows that the heating effect at constant power is roportional to frequency, which should thus be made as high as onvenient: in many materials additional advantage is obtained ecause ϵ'' increases with frequency. Dielectric heating equipnent is difficult to shield effectively and operation must usually e confined to assigned frequency bands where large amounts of tray radiation are permitted: the United States bands are given n Table 4. Electron tubes are available for these heating

Table 4

Assigned Frequencies for Dielectric Heating

Centre frequency, Gc/s	0·915	2·45	5·85	10·6	18
Deviation allowance, Mc/s	±25	±50	±75	±100	±150

ands,[79] the highest c.w. powers being provided by magnerons.[68, 69, 70] Typical outputs are 5 kW at 0·915 Gc/s and kW at 2·45 Gc/s but experimental tubes have given higher alues. The transmission-line output is either coaxial line or aveguide according to the frequency and power.

Some care must be devoted to the choice of a suitable operating frequency. For very-high-loss materials the field strength falls off rapidly with increasing penetration and thus the central portion of a large volume will be heated only slightly. The attenuation coefficient is given by

$$\alpha = \pi\epsilon''/(\lambda\sqrt{\epsilon'}) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

and, for example, in a material with $\epsilon' = 28$ and $\epsilon'' = 5\cdot6$ at 0·915 Gc/s, leads to the power density falling to 61% of the surface value in a distance of 2·5 cm.

Many materials to be heated by microwave energy are relatively low-loss dielectrics and, to secure efficient use of the available power, they must be placed in some form of resonant structure. The possible spatial variation of electric field imposes restrictions on the size and position of objects which must be uniformly heated. Thus in a TM_{010} mode circular cavity, the rate of heating is constant in an axial direction while more than 90% of its maximum value resides inside a radius of $0\cdot07\lambda$. With, for example, a plastic material of $\epsilon = 4$, heated at 0·915 Gc/s, the diameter must not exceed 2·3 cm to ensure this condition.

There is a number of important cases where it is required to heat non-uniform dielectrics in which a substantially uniform temperature rise is required, or where it is more economical to concentrate the heating as locally as possible. Rates of heating proportional to the heat capacity per unit volume, rather than uniform heating, are, of course, required to give a uniform temperature distribution. Since the permittivities and loss tangents of the materials comprising a composite or non-uniform dielectric load usually vary with frequency in different ways, it is sometimes possible to find a frequency giving a significant improvement in the uniformity of temperature rise. The relative rates of heating and temperature rise may also be controlled by selection of the mode of application of the electric field with, if necessary, relative rotation or translation of the dielectric sample.

(3.2) Particular Applications

Dielectric heating has been employed[59] for the bonding of plywood veneers. The glues should be of the thermal-setting plastic type and, where possible, selected for rapid polymerization at low temperatures. The coaxial system in Fig. 1(*a*) permits the electric field to pass through the glue face and wood in parallel, giving increased concentration of heating in the higher-permittivity glue. Frequencies giving a convenient size of microwave structure and rapid heating of the glue are in the range 1–6 Gc/s.

More bulky wooden articles can be bonded by employing,

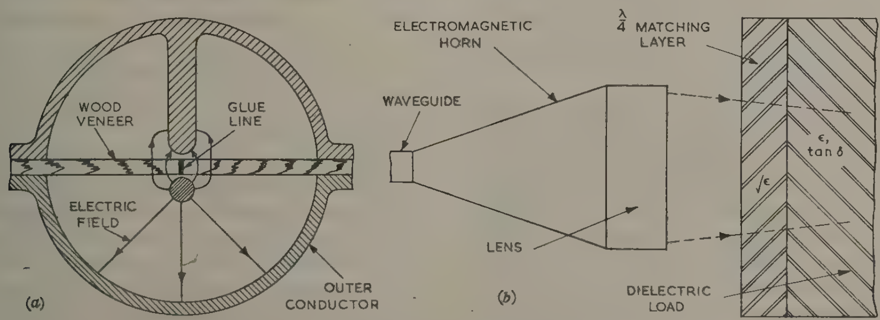

Fig. 1.—Methods of gluing wood by dielectric heating.

(*a*) Edgewise gluing of plywood veneers.
(*b*) Beam heating with focusing horn.

as shown in Fig. 1(*b*), an electromagnetic horn to beam the energy to the desired location. For normal-size articles, frequencies in the range 1·5 Gc/s are suitable. By means of a suitable reflector or lens, the beam may be made to converge to a minimum diameter of about half a wavelength. Thus, concentrated heating may be secured in internal portions of a load inaccessible for the application of electrodes. Surface reflections can be minimized by any of the usual matching techniques.

If the electrodes of a heating system are placed on opposite sides of a thin dielectric film, the electric field, and hence the rate of heating, is limited by breakdown of the unavoidable air-gap. In order to obtain the maximum electric field in the dielectric film for a given total power, flowing or stored, the spacing between the electrodes must be as close as possible to minimize the air-gap, giving an inconveniently low-impedance system for coupling power. By the use of fringe-field heating[36] shown in Fig. 2(*a*), the electric field may be made to spread in air giving a lower electric field there, to concentrate in the dielectric film and give a higher field, and to travel through a considerable

state. Too rapid heating must be avoided with frozen foods since the increase of loss tangent and dielectric constant o melting leads to intense local heating. Thawing and heatin times are reduced from the 15 min of conventional oven metho to about one minute. Extra heating of the surface for browning or crust formation may be secured by adjacent slabs[59] of hig loss material which attain much higher temperatures. Or method of determination of adequate cooking is to make th energy supplied proportional to the weight of food. Typic electronic ovens at 2·45 Gc/s provide powers of 0·8–1·6 k and the cooking operation involves placing the food in a glas china, plastic, paper or sometimes metal dish. Nutrition studies have shown that microwave cooking gives better retentio of vitamins and natural juices in the food.

(4) EXPOSURE TO INTENSE MICROWAVE RADIATION

(4.1) Tissues with Blood Vessels

Exposure of biological tissues to intense microwave radiatio leads to modification of their properties. Experiments[48] hav

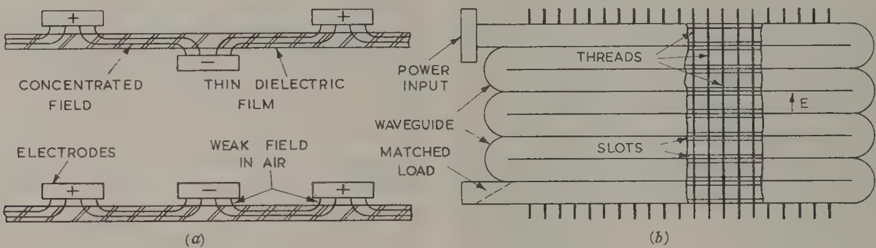

Fig. 2.—Dielectric heating of thin films and threads.

(*a*) Types of fringe-field heating for thin films.
(*b*) Waveguide heating of threads or strips.

length of film, thus giving a high impedance. Fig. 2(*b*) shows a waveguide, slit along the centre of the wide faces, and folded back on itself several times. Thread or strip material passing through the slits will thus lie in the position of maximum electric field. Energy propagated down the guide will be continuously attenuated but the average heating of each thread will be approximately the same. If the guide is sufficiently long only a small fraction of the input power reaches the far end, so that matching will be independent of load changes.

Microwave energy is a suitable medium for the heating and cooking of foods.[21] From this viewpoint the collection of foods comprising a typical meal is a dielectric load of low dielectric strength and variable permittivity and loss tangent. Fairly uniform heating is secured at frequencies of 1–3 Gc/s, owing to frequency variations of dielectric properties. At higher frequencies, the depth of penetration of the energy into the food is small, while at lower frequencies dielectric breakdown limits the rate at which power may be supplied.

At 3 Gc/s good uniformity of heating and matching with various combinations of food has been secured by the use of beamed radiation. Reflectors are used to direct the power onto the food from all sides to offset the effects of low penetration, and are arranged, as far as possible, to avoid returning energy towards the exciting horn that has not been dissipated by several passages through the food. The whole system may be screened in a metal case with a metal-gauze door for observation purposes. At 1 Gc/s, effective concentration of heating energy necessitates a resonator system.

To take full advantage of the rapid heating obtainable by dielectric methods the food should be in a partially pre-cooked

shown that the changes are probably due mainly to therm effects consequent upon absorption of the radiation. For t study of these changes, living tissues can be divided into tw classes[102]—normal tissue which contains blood vessels ar ischemic tissue which does not.

For example, practically any desired temperature can produced[46] in the various tissues of the thigh of a dog by proper controlling the output of the microwave generator and t duration of the exposure. The temperatures of the tissu begin to rise as soon as the microwave energy is directed towa them. The rise continues until approximately 20 min ha elapsed whereupon the temperatures begin to fall. Measur ments of blood flow in the thigh of the dog showed that marked increase in the flow occurred simultaneously with t fall in tissue temperatures: this increase acts as a heat-dissipati mechanism.[75]

Such effects have also been observed[38] in humans: the ter perature of the tissues again triggers a mechanism leading increased blood flow, which prevents the development of too hi temperatures. The mechanism of heat production in hum tissues by microwave radiation has been extensively examined[8] by means of pulsed sources in the 3 Gc/s frequency band. T results were analysed[16] and correlated with the respective p mittivities and thermal conductivities of the tissues involve Metals implanted in the tissue tend to set up standing wa with consequent[33] higher local rises of temperature.

Medical diathermy[47] involves controlled thermal effects whi are intended to be beneficial: the optimum frequency for norm tissues lies[92] in the u.h.f. region. The variations betwe measured and biologically effective dosages have been examine

ıd a comparison made[72] of the temperatures produced by icrowave and short-wave radiation. The blood changes in a t resulting from microwave diathermy have been studied.[51, 52] 'hen the temperature of the tissue exceeds a certain limit reversible changes take place: a simple example is the coagula-on of egg albumen.

The heat-exchange characteristics of animals are shown hematically in Fig. 3(*a*). Normal temperature is at the ordinate

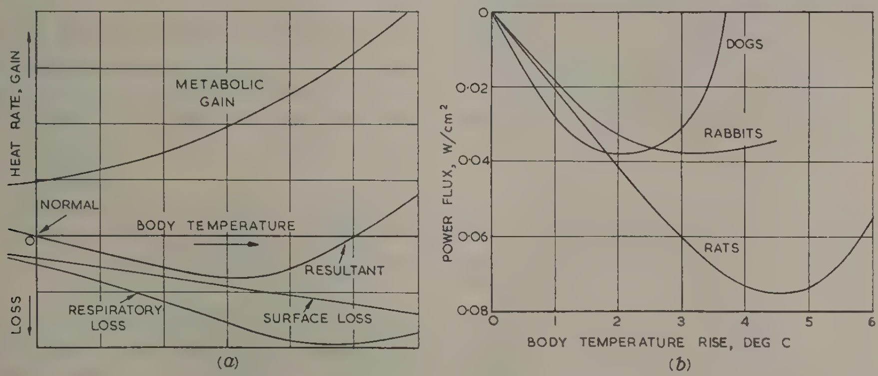

Fig. 3.—Heat-exchange characteristics of animals.

(*a*) Qualitative heat exchange in arbitrary units.
(*b*) Results on animals at 3 Gc/s frequency.

ıd increases to the right while heat gain from metabolism is ɔove the abscissa and heat loss below. The resultant curve ıows that at normal temperature the exchange mechanism aintains equilibrium, but for high fever temperatures the net ːchange becomes positive and eventually the animal dies. xperiments have been made at a frequency of 3 Gc/s on the eat-exchange characteristics of animals.[29] This involved study ' the absorption of electromagnetic energy, the ability of the ıimals to dissipate heat at elevated body temperature and the lation of field strength to body-temperature increases. Three ecies of animals of different sizes—rats, rabbits and dogs—were ːposed to a pulsed source of radiation via a horn antenna. ıe radiation was confined to a shielded enclosure which was ıed with absorbing material in order to give relatively free field ɔnditions. Fig. 3(*b*) shows the average results of the body-mperature rise above normal, plotted with respect to the icrowave power flux required to maintain this rise. Here icrowave power has been used as the additional source of heat ːcessary to cause the animal's net heat exchange to equal zero, other words, to maintain constant body temperature. The agnitude of the microwave power, then, represents the rate at ıich the animal would lose heat without this source, or his eat dissipation ability. The average absorption of each species ıs about 40% of the power in the animal's geometrical profile ıd the heat-dissipation ability was such that a flux density of 025 watt/cm² could be dissipated at a body temperature rise 1° C.

In other experiments,[58] mice were exposed to 10 Gc/s radiation th a power flux in the range 0·05–0·5 watts/cm². Their mperature rose at a rate proportional to the power flow and ath occurred in 50% of the subjects when a critical temperature 44·1° C was reached: this temperature is 6·7° C above the erage normal body value. The observed results were counted for by calorimetric considerations, the animals being sumed to be simultaneously heated and cooled.

(4.2) Tissues Without Blood Vessels

The absence of blood vessels in certain parts of the body makes them especially vulnerable to microwave irradiation: heat is then able to dissipate only by conduction to the surrounding vascular tissues. This is particularly true for the chambers of the eye and the hollow viscera such as the gall bladder, urinary bladder, and lumen of the gastro-intestinal tract. These areas are relatively avascular and largely devoid of effective mechanisms for regulating their temperature.

The effects of microwave irradiation on bone and bone marrow have been examined[32, 53] in both dogs and humans. Temporary or permanent sterility can result[54] from exposure of the testicular tissue to microwave energy. Damage to the reproductive tissue is to be viewed with particular concern as some geneticists believe that radiation far below the level which causes physio-logical damage may cause genetic damage that will not be apparent for several generations.[66] The changes in temperature in various avascular regions of an anaesthetized rabbit are given[48] in Table 5 as a function of time from the onset of irradia-

Table 5

Temperature Changes

Region	Temperature change for an irradiation time of					
	1 min	2 min	3 min	10 min	20 min	30 min
	deg C	deg C	deg C	deg C	deg C	deg C
Ileum	+4·2		+14·4	+29·5	+38·5	+42·9
Stomach ..	+1·8	+3·4	+5·4		+19·2	+23·1
Gall bladder ..		+0·1	+0·3	+1·8	+4·0	+6·3
Urinary bladder	+1·3	+2·1	+3·0	+5·6		+9·7
Rectal	+0·1		−0·1	+0·2		+0·8
Oral	−0·2		−0·2	−0·5	−0·9	−1·2

tion at a frequency of 2·45 Gc/s. These findings also point to the limitations of the oral and rectal temperatures, which can be seen to have remained almost constant, as indications of injurious temperatures in the avascular regions. Exposure of the head alone resulted in the temperature of the brain rising by 6° C with fatal consequences.

The demonstration of the production of opacities in the eyes of animals,[24, 26, 74] as a result of prolonged exposure at high field intensity, represents an important biological effect of microwave radiation. Although experiments have been carried out[76] at 10 Gc/s, most work has employed standard diathermy equipment operating at a frequency of 2·45 Gc/s. Determina-tions have been made of the effect of irradiation on the tem-

peratures of the orbital tissues and the aqueous and vitreous humours in dogs and rabbits.[27, 100] In most cases the actual temperatures of both humours, after exposure, were consistently higher than that of the deep orbital tissues. The aqueous and vitreous humours are entirely avascular and the rapid rate of cooling observed was shown to be due to blood circulating in the adjacent vascular tunics.

Rabbits have been used[46, 78, 103, 104] for most experimental work since their eyes are very nearly the same size and shape as those of humans. In one case[12] a cataract, a form of white cloudy growth, developed after a 10 min exposure to about 100 watts at 2·45 Gc/s. Results of temperature measurements within the eyeball at two different microwave frequencies are shown in Fig. 4(*a*). It will be observed that at 2·45 Gc/s the highest

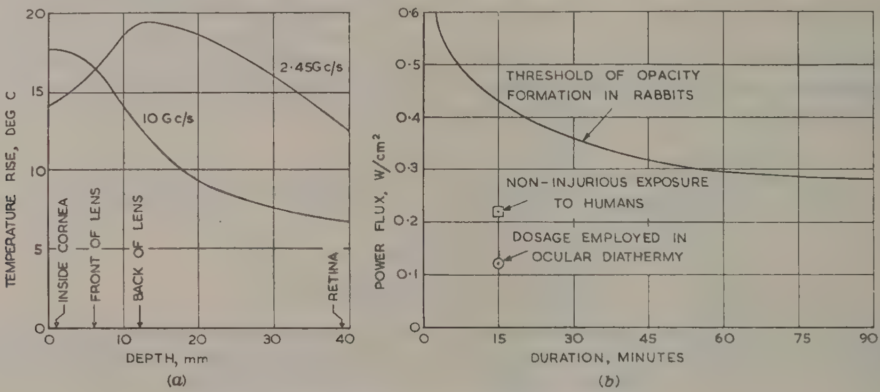

Fig. 4.—Effects of microwave irradiation of the eye.

(*a*) Temperature distribution.
(*b*) Production of opacities at 2·45 Gc/s.

temperature occurs near the back surface of the lens which itself consists of protein very easily damaged by heat. This plot explains the cataracts found at 2·45 Gc/s and the opacities formed[76] at 10 Gc/s inside the cornea and on the anterior segment of the lens.

In another series of experiments[101] the durations and power fluxes of the single exposures were varied between the limits of 3·5 min at 0·59 watt/cm^2 and 90 min at 0·24 watt/cm^2. The damage observed was mainly in the lens of the eye and was classified into three groups:

(*a*) Minimal opacities such as small black dots not detrimental to vision.
(*b*) Circumscribed opacities, dense enough to interfere with vision, consisting of diffuse cores with dark ray-like borders.
(*c*) Large opacities which obstructed vision, comprising large areas with peripheral crescents, dense linear projections and vacuoles.

All the lens opacities developed within 14 days after irradiation, the degree of damage being inversely proportional to time of appearance. The results were expressed as in Fig. 4(*b*), the solid line drawn through the experimental points representing an empirical limit above which production of lens opacities is a near certainty. The extreme susceptibility of the eye to a power flux greater than 0·4 watt/cm^2 is indicated by the precipitous slope in this region. Also shown are exposures which were not found detrimental to human eyes and a standard ocular diathermy treatment.[13]

Several investigators[24, 26, 48, 74, 103] have detected opacities within 2–42 days of multiple exposure to microwave radiation. More recent experiments,[105] in which care was taken to ensure that any one of the exposures was below the threshold value, appear to confirm a cumulative effect. Pulsed irradiation[76, 105] seems to show an enhanced tendency to the formation of opacities as compared with that due to continuous sources o the same average power. Microwave irradiation is also found to reduce the activity of certain enzyme systems in the eye. N opacities have been observed at the lower frequencies o 0·2–0·5 Gc/s even with whole-body exposures near the leth level.[106] From these various results it cannot yet be conclud with certainty that thermal effects alone are the cause of opaci formation.

(5) OPERATIONAL HAZARDS TO PERSONNEL

(5.1) Exposure Investigations

There is some uncertainty as to what constitutes a dangero microwave radiation field.[9, 99] Damage to a biological syste by some hazard can usually be expressed[50] by a relation the form

$$Ct = K \qquad \ldots\ldots \quad ($$

in which C is the concentration and t is time. K is a numb which, when it exceeds a certain characteristic value, alwa means damage to the organism. Conversely, there is no dama when it is lower than this characteristic value. The value abo which damage is always observed, and below which damage not to be expected, is often called the threshold dose. A important factor in influencing tissue damage is the varyi ability of a biological system to repair itself or to restore a degr of its disturbed function towards normality while still under t influence of the damaging agent. Not only do various tissu have different capacities in this regard but the same tissue different individuals will vary in its recovery capacity. Furth the same tissue in the same organism will have a differe response at different times. Although eqn. (3) thus applies wi varying degrees of accuracy, it forms a useful concept and t values of K in different cases must be sought.

The concentration C is related to the field intensity in the fr space occupied by the organism, the ability of the various tissu to absorb the energy[34] and the depth of the tissue beneath t surface of the organism. The absorption of electromagne energy[88] by various human tissues in the range 0·15–10 Gc can be represented by the triple arrangement of skin, su cutaneous fat and deep muscle tissue shown in Fig. 5(*a*). Sin the depth of penetration has been shown[89] to be sufficient small, the deep tissue layer may be assumed to extend to infinit Using known measured values[88] for the relative permittivity a loss tangent of skin, fat and muscle, the proportion and distrib tion of absorbed energy in the respective parts of the body ca

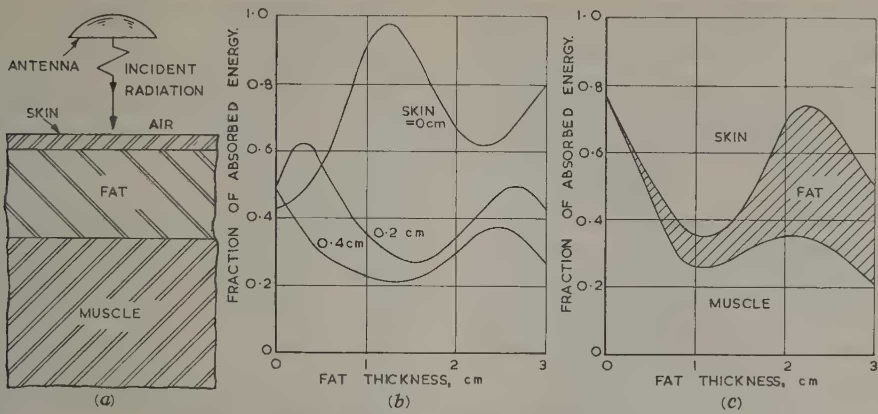

Fig. 5.—Absorption of microwave energy by the body.

(*a*) Triple-layer arrangement.
(*b*) Absorption of energy by the body at 3 Gc/s.
(*c*) Distribution of absorbed energy at 3 Gc/s for a skin thickness of 0·2 cm.

›e calculated,[90, 91] Fig. 5(*b*) gives the fraction of airborne adiation at 3 Gc/s absorbed by the body as a function of thick-ıess of subcutaneous-fat layer: the parameter of the different ːurves is the assumed skin thickness in centimetres. At 0·15 Gc/s he penetration is much deeper while at 10 Gc/s most of the nergy is absorbed in and near the skin. The distribution of ıbsorbed energy is shown in Fig. 5(*c*) for a frequency of 3 Gc/s ınd skin thickness of 0·2 cm. Results for greater thickness eveal, as expected, increased absorption in the skin with decrease n the deeper tissues.

Attempts have been made on the basis of experimental results o give limits to the factor C in eqn. (3) for prolonged dosage. The safe values arrived at apply only to thermal effects due to nean powers and are related to a rise of 1° C in body tempera-ures. Limits have been provisionally given[87] as 0·03 watt/cm^2 or frequencies below 0·5 Gc/s where deep heating exists, ·01 watt/cm^2 for 0·5–3·0 Gc/s where complicated deep and urface effects occur and 0·22 watt/cm^2 at frequencies higher han 3 Gc/s where sensory perception of the heat absorbed in he skin provides some safeguard. Above 0·1 watt/cm^2 the hort-term body effect of heat stroke becomes predominant, the igher temperatures obtaining having fatal results.[63]

Eye effects are extremely important but rather complicated to nterpret. At low levels there is a long-term integration result-ng, as has been noted, in the formation of opacities and ataracts. A field appropriate to 0·01 watt/cm^2 would require many years to cause harmful effects. Owing to their small size and lack of blood vessels, the eyes have only a short time-constant and a power flux exceeding 0·5 watt/cm^2 can soon cause damage. This information is based mainly on experiments on animals. The formation of bilateral cataracts has been reported[49] in a technician who was exposed in daily work to an average level at 1·7–3·4 Gc/s of 0·005 watt/cm^2, with intermittent exposure during the preceding three days to 0·12 watt/cm^2 for a total of 2 h. In general, the available data[1] on human cases is too meagre for a statistical analysis to be made.

(5.2) Instrumentation

The safe limits of field strength at various microwave frequencies having been decided, it is necessary that any particular equipment should comply with them. Much can be achieved by calculation from data relating to the equipment. A typical antenna is a paraboloid fed by a horn which has a gain of 8 dB, the field at the edge of the reflector being 10 dB less than that at the centre. The field outside the reflector will be less than this, and thus, except in the region between the horn and the reflector, the direct field from the horn is less than it would be from an isotropic radiator. For example, with a radiated power of 10 kW the minimum safe distance is of the order of 3 m and thus the danger area due to direct radiation from the feed is small.

The field due to the main beam is of greater importance and Fig. 6 gives[94] the values for distances R to $10R$, where R is the

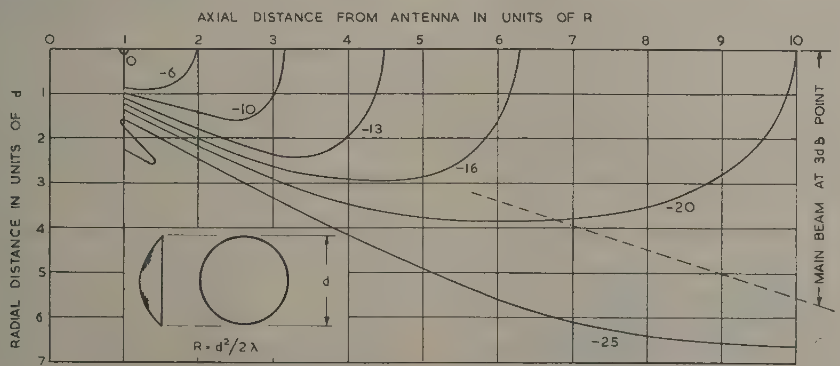

Fig. 6.—Power-flux contours of a radiating paraboloid.

Parameter is decibels relative to $1 \cdot 8P/d^2$, where P is the input power to the antenna.

Rayleigh distance of the antenna. The contours are drawn relative to the maximum power flux which is taken as the flux on the axis at distance R. This maximum value is approximately $1 \cdot 9P/d^2$ but the actual flux will exceed this by up to $3\frac{1}{2}$ dB at some positions within the Rayleigh distance. The values computed are those in free space, and perfect ground reflection will enhance the field strength at some points by 6 dB.

Reasonable parameters for a tropospheric-scatter system on a frequency of 0·855 Gc/s would be $P = 20$ kW and $d = 10$ m, giving $R = 1 \cdot 41 \times 10^2$ m and a maximum power flux of 0·046 watt/cm². Taking a safe flux of 0·01 watt/cm² and including a safety factor of 4, the safe contour is −13 dB. If the mean height of the antenna is 10 m above a level ground, the whole of the ground between 270 m and 620 m in front of the antenna is within the danger zone since the flux exceeds the safe value at heights below 2 m. This unsatisfactory position can be remedied either by increasing the mean height of the antenna to 14 m or by tilting the antenna upwards by 0·45°, i.e. about one-fifth of a beam width. Such an antenna could therefore safely be placed on a cliff overlooking the sea, or with ground sloping away in front of it at an angle exceeding 0·45°.

The field at distances from the antenna less than the Rayleigh distance is largely confined within a cylinder the base of which is the reflector. The field below this cylinder is generally about the same as it is at the same height at the Rayleigh distance. For the antenna considered above, the field at a height of 2 m above the ground varies with distance from the antenna but is always more than 20 dB less than the maximum power flux up to a distance of 200 m. The field increases to about 10 dB at 340 m and then decreases continuously, reaching 14 dB at 680 m and 20 dB at 1 380 m.

Even medium powers and small antennae can be dangerous: for example, 100 watts radiated by a 1 m diameter paraboloid gives a maximum flux of 0·023 watt/cm², well above the tolerable limit. The Rayleigh distance is, however, quite small, being only 5 m for a frequency of 3 Gc/s. There is little information on the effect of high fields lasting for a microsecond or so, and it is usually assumed[96] that the relevant power flux from a pulsed radar equipment is the power averaged over one repetition period. The average flux at any given location is further reduced if the antenna is scanning. These power fluxes for microwave equipment can be compared with the value of 0·14 watt/cm² due to the energy of sunlight reaching the earth.

Many sites are prone to effects such as ground reflections, and some form of monitoring or instrumentation is then necessary. Pending the complete elucidation of the exact nature and biological effects of microwave energy, it might be expedient to undertake a controlled monitoring programme employing suitable biological subjects.

Satisfactory monitoring of power flux would be facilitated by a simple instrument which indicates whether or not the s value is being exceeded. Neon tubes are qualitatively use but tend to be erratic and, quantitatively, may be quite m leading. The field strength at any point near the antenna c be measured with a standard-gain horn and calibrated receiv Such a device should be broadband in performance so tha wide spectrum can be covered with only a few instrumen Power-measuring devices of the bolometer type have be employed for personal monitors of a portable nature.

Another approach to the problem of monitoring is the use biological simulants. For example, the effect of exposure on t eye can be examined by noting the response of gelatin mixtu Studies have been made[50] of the temperature distribution ins spheres, a concentration of 30% being chosen to represent t protein content of most mammalian eyes. Temperatures various depths were measured with thermistor beads and t values found at 10 Gc/s with a power flux of 0·015 watt/c agreed closely with those obtained under similar conditions the eye of a cow. Some improvement was achieved by loadi the simulant with absorbing material.

(5.3) Methods of Protection

The maximum safe power flux for personnel has been co sidered by several authorities[23, 60, 61, 107] and, in general, the val of 0·01 watt/cm² over the whole frequency range has be provisionally accepted. It is evident that areas in which such power flux is exceeded must be fenced in or otherwise enclos to prevent unauthorized or accidental entry by civilians a workers who may be in the vicinity. For personnel who m work in hazardous conditions a code of practice is desirable.[10] This should explain the dangers present, give information minimum safe distances, request personnel to be on their gua for symptoms of heat or discomfort and give details of protecti arrangements. First-aid measures should include artific respiration, oxygen administration and means for rapid cooli of the body. Normal clothing absorbs microwave radiati and thus gives some protection to the body although tending aggravate the problem of keeping cool: thus, overalls made reflecting material would appear advantageous.

It is essential that the eyes be protected, and measurements ha been made[28] on materials which are optically transparent b provide microwave shielding, with a view to their use in fa shields or goggles. The power-transmission factors given Table 6 represents both reflection and absorption of microwa energy, the d.c. surface resistivities quoted being an indicati of the former. The results show that good electrical conductiv is essential for microwave shielding, whereas this characteris is not generally compatible with light transmission and oth psychological and physiological factors to be considered in t application to protective goggles. The attenuation of t

Table 6

SHIELDING PROPERTIES OF VARIOUS MATERIALS

Material	Power-transmission factor at			
	5·9 Gc/s	9·7 Gc/s	18·8 Gc/s	0·55 micron
	%	%	%	%
Gold film, 11 mμ thick on plastic (300 ohms/square)	23	10	0·8	49
Gold film, 30 mμ thick on plastic (12 ohms/square)	0·16	0·1	0·01	24
Gold film, 75 mμ thick on glass (1·5 ohms/square)	0·04	0·01	0·004	3·2
Corning glass, 1·5μ conductive coating (15 ohms/square)	1·6	1·2	0·08	45
Electraplane glass, 300 mμ conductive coating (70 ohms/square)	9	10	8	80
Copper mesh (20 per inch)	0·1	0·2	0·2	50
Copper mesh (8 per inch)	1·0	1·3	2·5	60

etallic films depends critically on the thickness, while the ielding improves as the frequency increases. A pair of suitable oggles would have a gold film on the lenses and wire mesh on e sides where its undersirable viewing properties would not atter. The 3·2% transmission is not too low for outdoor work nd is permissible for indoor work in well-lighted rooms, proided that the greenish-blue tinge of the gold film can be tolerated. or particular instrumental applications where visual acuity is ss important, wire mesh or the Corning heating-panel glass ay be more suitable. Exposure to direct microwave radiation ay be avoided by the remote viewing of energy channels with, r example, telescopes, periscopes or closed-circuit television.

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A DIGITAL COMPUTER STORE WITH VERY SHORT READ TIME

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SUMMARY

The paper describes the principles of operation and the construction f storage units with a very short access time for reading. One form f the store which has been constructed has a capacity of 200 000 bits f permanent information and another has been built with a capacity f 100 000 bits, the whole contents of which may be changed in under ne minute. The technique employed permits the construction of very rge stores at low cost. Each digit cell is formed, basically, by two ts of windings which form the primary and secondary of a transrmer and the two binary states are determined by the presence or bsence of a piece of linear ferrite material coupling the windings. he access time is largely determined by the physical size of the store d the speed of operation of the associated circuits; a time of 00 millimicrosec is typical.

(1) INTRODUCTION

It is well known that a store with rapid-reading access is of onsiderable value in high-speed digital computers. Several ethods of achieving this have been devised, including punched-ard reading devices[1] and wired systems of square-loop magnetic ores.[2,3] With the present development of very-high-speed omputing circuits using transistors, the access time to the omputer storage devices may be the major factor in determining he overall computer speed. It is therefore necessary to improve hese access times, and in particular the access time of that ection of the store which is retaining permanent or semi-ermanent information.

Magnetic rather than photo-electric devices seem to offer the ost satisfactory solution to this problem, especially if advantage taken of the fact that square-loop cores with their inherent witching time are unnecessary. Alternatively, it is realized that store could be constructed in which the two binary states are epresented by the presence or absence of resistive or capacitive lements. In this case the presence of the element is detected by easuring the current which flows in a read-out wire when a oltage is applied to a drive wire, using a current amplifier with ery low input impedance. This has not been done, because hanges of stored information, when desired, are not so con-venient as with the magnetic elements.

The storage system to be described can be constructed in a umber of different forms determined by the use for which it is tended. In the paper, two forms are chosen: first, a permanent orm of 200 000 bits, 'permanent' being interpreted as permitting ery occasional changes by a simple manual operation; secondly, semi-permanent form of 100 000 bits permitting partial or omplete change of the stored information in one minute. The ermanent form allows slight modifications or additions to be ade during the lifetime of a computer, whilst the semi-ermanent form allows severe changes between successive roblems.

A special feature of the storage system is the low cost per stored it, and it is estimated to be between one-tenth and one-thirtieth of the cost per bit for the square-loop magnetic-core matrix store.

The system has a number of advantages over the permanent store produced by special wiring of square-loop cores;[2,3] the construction is simpler and the stored information may be changed, the read-out time is less and the cost is lower.

(2) PRINCIPLES OF OPERATION

The storage system makes use of the presence of a rod of magnetic material to represent a stored 'one' digit. The rod is positioned to couple the primary and secondary windings of a transformer. In the 'zero' state, the ferrite rod is absent and the coupling between primary and secondary windings is very small. The state of the storage cell is determined by passing a current pulse through the primary winding and observing whether there is an e.m.f. induced in the secondary. In contrast with storage systems employing square-loop magnetic cores, which have fixed minimum switching times, this system makes use of linear ferrite magnetic material and there is no inherent speed limitation.

Access to information in the store is obtained by reading out the digits of a word in parallel. The general arrangement of the store is shown in Fig. 1. The primary windings are con-

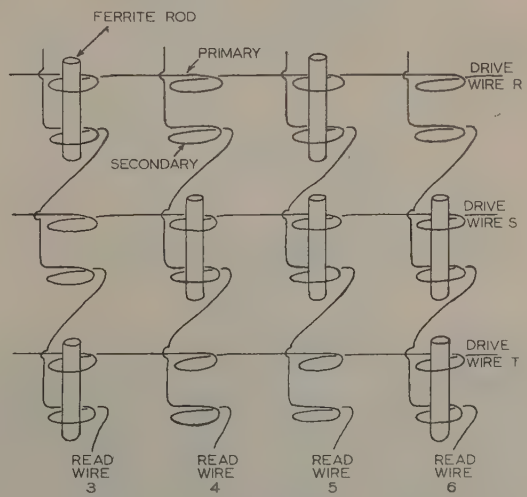

Fig. 1.—General arrangement of storage system.

nected in series to form the horizontal drive wires and the secondaries are similarly connected to form the vertical read wires. When a current pulse is passed through one particular drive wire, S, an e.m.f. will be induced in read wires 4, 5, and 6. In contrast, a negligibly small e.m.f. will be induced in read wire 3 because there is no ferrite rod linking the windings of the transformer formed at the intersection of drive wire S and read wire 3. The signals from read wires 4, 5, and 6 indicate

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'one' digits, and the absence of a substantial signal from read wire 3 indicates a 'zero' digit. The individual words of the store are selected by passing currents through different drive wires. The waveform of the induced e.m.f. is ideally the time differential of the drive current and typical waveforms are shown in Fig. 2.

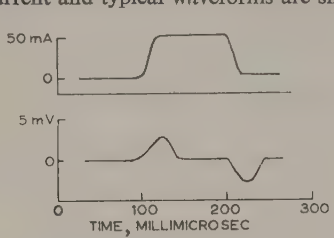


Fig. 2.—Store waveforms.
Upper trace: drive current waveform.
Lower trace: output from secondary of read-wire transformer.

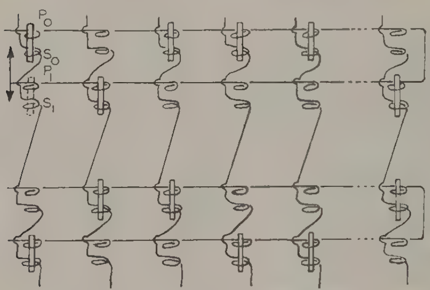


Fig. 3.—Double-sided system.

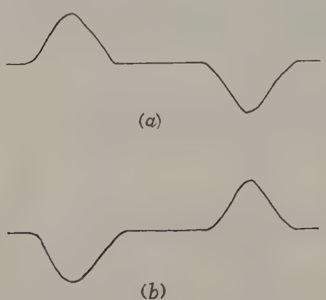


Fig. 4.—Double-sided system waveforms.
(*a*) 'Zero' output signal.
(*b*) 'One' output signal.

In an alternative arrangement (Fig. 3) two sets of windings are provided for each digit cell and the magnetic core has one of two possible positions. In the upper position, the cell is in the 'zero' state, windings P_0 and S_0 are coupled and the waveform of the e.m.f. induced in the read wire is as shown in Fig. 4(*a*). With the core in the lower position, coupling the windings P_1 and S_1, the induced e.m.f. is in the opposite sense, Fig. 4(*b*), because the primary windings are connected so that current flows through them in opposite directions. If only the first pulse of the pair appearing on the read wire is observed, a positive read-out signal is obtained for a stored 'zero' digit, and a negative signal for a 'one'. This arrangement gives a greater discrimination than the single-sided system.

(3) THE CONSTRUCTION OF THE STORE

The mode of construction of the store is important because it has considerable influence on the electrical design and operation of the system. Furthermore, by special attention to tl manner of construction the overall cost may be considerab reduced. The basic requirements for the system are a lar number of primary and secondary windings suitably connecte with arrangements for inserting or moving pellets of magnet material. A typical size of store required is 4096 words, approximately 200 000 bits, so that the technique employed fi constructing individual digit cells must be easily repeatabl One method investigated makes use of printed-circuit technique A pattern of conductors as shown in Fig. 5(*a*) is produced

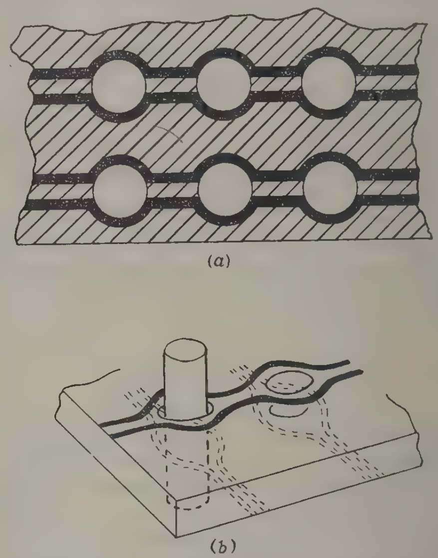


Fig. 5.—Store construction using printed circuits.
(*a*) Pattern of conductors.
(*b*) Arrangement of primary and secondary windings.

cover an area of 7 in × 7 in containing 2916 digit cells. O printed-circuit board with this pattern is used for a set of prima windings and a second board mounted beneath the first ar cemented to it, with the conductors at right angles to those c the first board, serves as a set of secondary windings. Hol are drilled for each digit position [Fig. 5(*b*)] and ferrite roc 1·6 mm diameter and 6 mm long, are inserted where require This method of construction, whilst being suitable for sma capacity stores, appears unsuitable for large versions. The co of producing the holes, by drilling or punching, is excessiv and the maximum size of the boards on which printed wirir may be produced is limited; this results in a large number connections having to be made between the boards.

A preferred system is to use continuous conductors over tl extent of the store. The scheme adopted employs a wov mesh of tough enamel-covered wire which can be readily man factured. The primary or drive wires are formed by adjace pairs of wires of the warp, and the read wires are formed l adjacent pairs of the weft (Fig. 6). The mesh is woven wi 33 s.w.g. wire, and there are approximately 18 meshes per inc The mesh is backed with a thin layer of Plasticine, and ferri rods 1 mm diameter and 6 mm long are inserted in the me where required and held by the backing material. In the Figur positions A, B, C and D are information cells, and rods may l inserted in these positions to represent stored 'one' digit Other ferrite rods are placed permanently in surroundir positions. These additional rods act as 'keepers' and provic return paths for the flux in the information rods. In tl absence of these keeper rods, the situation depicted in Fig.

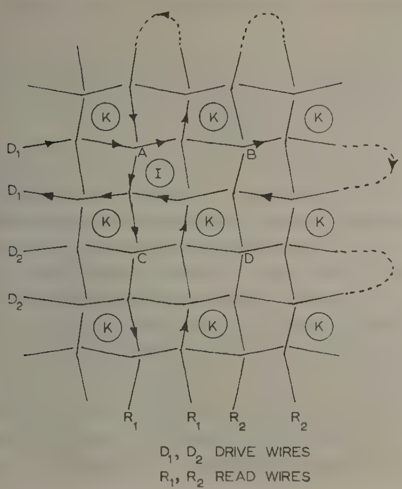


Fig. 6.—Store using woven mesh.

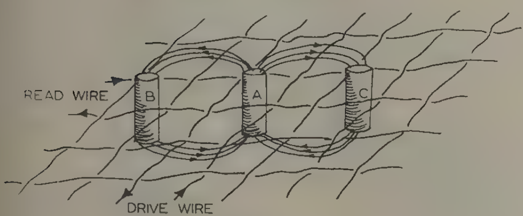
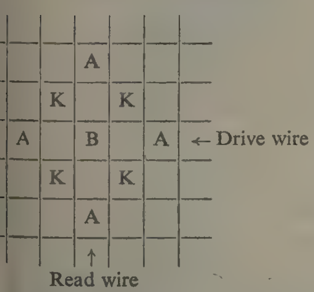


Fig. 7.—Situation with no keepers.

ises. If a pulse of current passes through the drive wire own, the flux change in the ferrite rod A will induce an e.m.f. the read wire. The rods in positions B and C will form turn paths for the flux. The change of flux in rods B and C duces e.m.f.'s in the read wire in the opposite sense, and may most cancel the read signal from rod A. This cancellation ect may be reduced if the rods are spaced sufficiently far art, but since close packing is required, keeper rods are sential. The improved magnetic circuit also results in an creased read signal. Results of tests are given in Table 1.

Table 1

READ-OUT AMPLITUDE FROM MESH STORE



Rods in position	Peak amplifier output mV
No rods	50
* B	1 000
* A	−240
* A, B	300
K	80
K, B	1 500
K, A	60
K, A, B	1 300

K, Keepers.
A, B, Information rods.

* These cases are not permissible.

is shows the magnitude of the peak read signal under various nditions. Under normal working conditions with keeper rods sent the 'zero' signal is less than one-sixteenth of the 'one' nal.

The woven structure has the advantage of low cost and has the windings conveniently disposed around the holes for the rods. As it is possible to obtain large sheets of the material, the problem of making joins between small sections does not arise. It has the disadvantage of having a small but finite capacitance between read and drive wires. Alternative modes of construction have been devised in which the read wires are spaced from the drive wires in a way similar to that in the printed-circuit arrangement. Such schemes generally require accurate alignment, since the medium in which the one set of wires are held must have holes through which the rods can pass, and these holes must register with the corresponding holes for the other set of wires.

The general arrangement of a store for 4 096 words each of 52 bits is shown in Fig. 8. The read wires are arranged in 52 sets

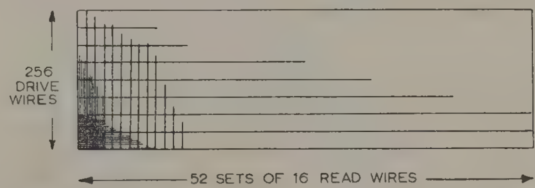


Fig. 8.—General arrangement of woven-mesh store.

of 16. A read selection system chooses one word out of the 16 which are all driven at once. All the first digits of each word are grouped together, as are the second, third and remaining digits, to simplify the read selection arrangement and keep leads short. The read-wire selection system is shown in Fig. 9. The

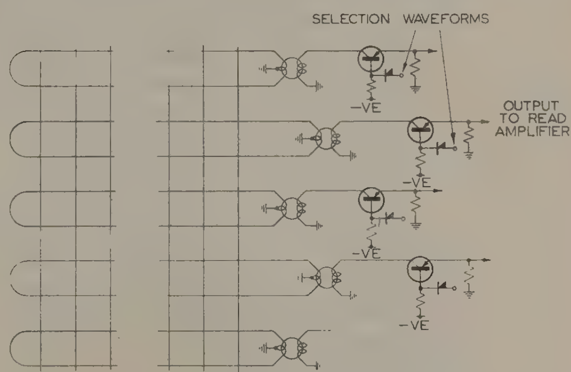


Fig. 9.—Read-wire selection.

two wires of each pair of read wires are joined together at one end and are connected to the primary of a 1 : 1 transformer at the other end. The secondary winding feeds a transistor switch. The collector is connected to one end of the secondary and the base is connected via a high resistance to a negative supply. In the open position the transistor is kept in the 'off' position by a diode which holds the base positive with respect to both the emitter and collector. In the selected position, the diode is released and sufficient base current flows to drive the transistor into the saturated region. There is then a low-impedance path between emitter and collector with a negligible transmission time. Using junction transistors with an α cut-off frequency of 10 Mc/s it is possible to complete the switching operation, changing from one of the 16 blocks to another, in under 1 microsec, allowing for the decay of the switching transient.

Only one out of each group of 16 transistors is on at any one time, so that there are a total of 52 transistors on together, each one feeding into a read amplifier. The transformers are employed to reduce the signals resulting from capacitive pick-up between drive and read wires.

Each of the 256 drive wires is supplied by its own drive transistor via a transformer, each transistor driving 832 digit cells. The current pulse is produced in a drive line by turning on the transistor driving it into the bottomed state. The transistor operates as a switch and the current-pulse amplitude of 50 mA is defined by the supply voltage and the circuit impedance. The pairs of drive wires have distributed inductance and capacitance and effectively form a transmission line with a characteristic impedance of 100 ohms. The time delay is 4 millimicrosec/ft, and correct matching is desirable because the read output waveform depends on the shape and rate of rise of the drive pulse. Accordingly, resistances of 100 ohms are connected across the ends of each drive line pair. The above figures for characteristic impedance and delay time are for the mesh without the information rods but with the keepers. The change in these values when all the information rods are inserted is small. With this arrangement of drive transistors any one of a block of 256 words may be selected by turning on the appropriate drive transistor. Switching between the 16 blocks is accomplished by the read selection circuits. An alternative driving system is described in Section 4.

The amplitude and shape of the pulses at the output of the read amplifier are controlled by a number of factors. The drive waveform [Fig. 10(*a*)], which gives a current pulse of

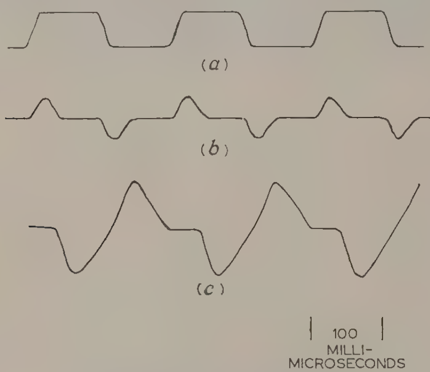


Fig. 10.—Store waveforms.

(*a*) Drive-current waveform.
(*b*) Waveform at secondary of read transformer.
(*c*) Read amplifier output.

50 mA, has a rise time of 20 millimicrosec. The corresponding read signal at the secondary of the read transformer [Fig. 10(*b*)] is the differential of the drive waveform, but is delayed by the transmission time in the drive and read wires of the store. The read amplifier, which uses three drift transistors and provides output pulses of 1 volt peak, has a response extending to 5 Mc/s. In consequence, the read pulses are broadened and delayed when they emerge from the output of the amplifier [Fig. 10(*c*)]. This results in an access time of 100 millimicrosec and a cycle time of 200 millimicrosec, which is adequate for the required application. A reduction by a factor of at least two is possible if an amplifier with greater bandwidth is employed.

The mesh has a width of 54 in, this being the length of the read wires, and the length of the mesh and drive wires is 8 ft. These dimensions include spaces left between blocks of digits. To economize in space the mesh is folded back on itself halfway along its length. The mesh is mounted on a board which carries the Plasticine backing material and is mounted with t drive wires vertical. The store is loaded with the aid of a speci jig which has a number of holes to correspond with the requir apertures of the mesh. The jig is loaded with information a keeper rods, and by placing it in the correct position on t mesh, these rods may be inserted.

(4) THE CHANGEABLE VERSION OF THE STORE

The permanent store is of value for holding standard libra sub-routines. An alternative version of the store has been ma in which stored information is changeable. This may be us for some of the more extensive programmes and for data whi are not changed during the course of a calculation. This versi of the store holds over 100 000 bits and the entire contents, part of it, may be changed in under one minute.

The store is constructed using the nylon mouldings shown Fig. 11. These are assembled in blocks of 32. Each mouldi has 26 tubes which contain ferrite rods 1 mm diameter and 3 m long, half the length of the tubes. The blocks are assembled

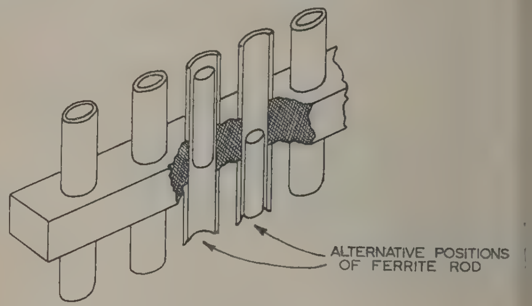


Fig. 11.—Nylon moulding.

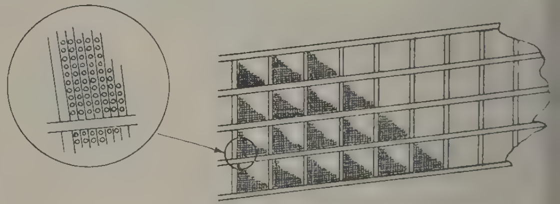


Fig. 12.—Nylon mouldings assembled in frame.

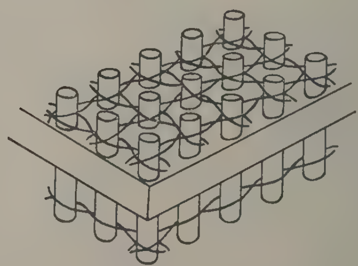


Fig. 13.—Wiring of changeable store.

a frame (Fig. 12), 32 blocks giving a storage capacity 32 × 32 × 26 bits, or 512 words of 52 bits. The complete st comprises four frames. The balanced system of windings is us the ferrite rods being able to occupy positions at one end of tubes or the other. The wires are threaded in a zigzag man between the tubes starting at one end on the first side of the fra (Fig. 13). When the wire has traversed one length of the f

de, the frame is turned over and the wire is threaded between e tube projections on the opposite side. At the end of the cond traverse, the wire is threaded back, completing the turns the opposite side. Finally the fourth traverse completes the rns on the first side. The read wires are wound in the same ay, but at right angles to the drive wires. The read wires are ound in the reverse sense on the opposite side so that the read gnal will change sign as the ferrite rod moves from one end of e tube to the other.

The information in the store is changed by blowing the rods om one end of the tube to the other, using a row of air jets which avels along the length of the store. This method is preferred to possible mechanical system in which the ferrite rods are moved pusher rods. With this there is the danger of damage to the ouldings or to the wiring in the event of incorrect registration. here is also an alternative electromagnetic system in which the ds are moved by a travelling row of electromagnets, but this considered undesirable because the rods have a high stiction the tubes and there is a danger that a rod in an adjacent row ay be moved unintentionally because a very strong field must applied.

The store is arranged with two frames in line with a further o frames placed behind at a distance of 4 in. The rows of ts are mounted on an arm which travels along a rigid girder the top of the frames. The travelling arm has 104 jets at one de of the frame and an equal number on the reverse side ig. 14). The equipment is duplicated for the second pair of

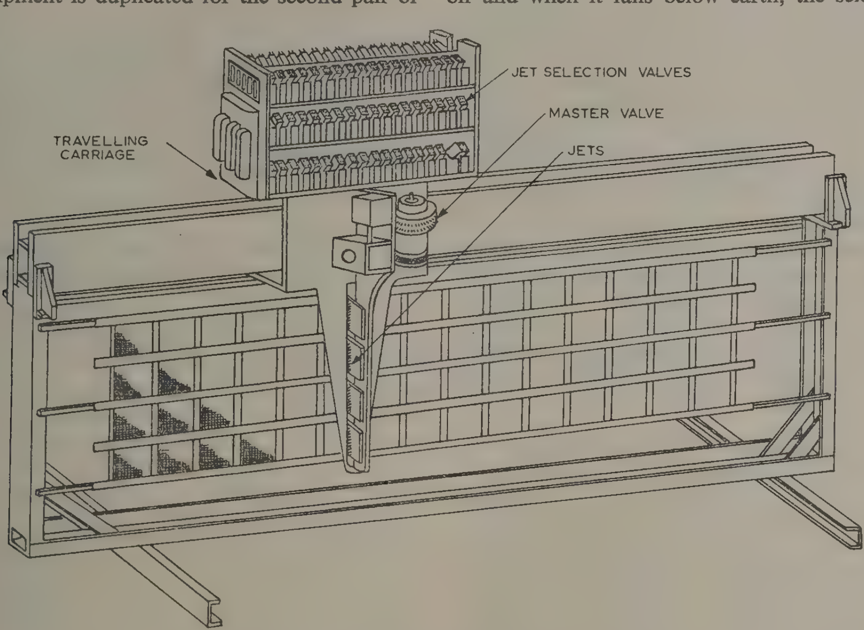


Fig. 14.—General arrangement of changeable store.

ames. The jets are grouped in pairs, one on each side of the ame, and are supplied by a change-over valve which admits air one or the other jet. The change-over valves are controlled small solenoids, and the complete group of 104 valves is first according to the information to be stored. Then, at the quired time, a pulse of air is supplied to all the valves and the ds are blown to one end of the tubes or the other. The air pply is controlled by a master valve which is operated by a oto-electric sensing device. In addition to the normal tubes in mouldings, a special tube is provided with a smaller hole. light source is carried on one side of the frame on the travelling arm and a photo-transistor on the other. As the registration hole is smaller than those in which the rods move, the master valve is operated only when the jets are correctly aligned in the centre of the holes. The air is supplied from a reservoir at a pressure of 10 lb/in^2. All interconnecting pipes are kept as short as possible to reduce the time of the air pulses to reach the jets. The time allowed to move a row of 104 pellets is 100 millisec and the complete store can be changed in 55 sec.

The loading operation will be under the control of a special programme on the computer to which the store is to be connected. Since the reading time is negligible compared with the setting time, it is possible to check that the rods have moved correctly each time they are set. Facilities are available for reversing the jet carriage to correct a setting error. The tubes are sealed by a thin non-magnetic wire cemented across the ends to prevent the rods falling out. The wire is cemented to every tube to prevent the possibility of a shock wave travelling along it and thus moving a rod unintentionally.

An alternative and more economical drive-selection system employs the diode matrix shown in Fig. 15. The rows of the matrix are controlled by transistors connected as emitter-followers and the columns are controlled by transistors which operate as switches. The transistors connected to the columns are controlled by waveforms applied at A_1, A_2, A_3, etc., which swing between levels above and below earth potential. When the waveform is above earth potential, the transistors are cut off and when it falls below earth, the selected condition, the particular transistor bottoms and the collector is held at approximately earth level. The controlling waveform for the emitter-followers applied at B_1, B_2, B_3, etc., swings between just above earth potential and -9 volts. The drive lines of the store are connected between the points PP, QQ, RR, etc. To drive a particular line ZZ, the level of A_3 is taken negative and B_3 is taken to -9 volts. A potential difference of 9 volts is then established across the line ZZ. The drive lines are terminated, as before, in their characteristic impedance. Certain other drive wires in the store experience a change in potential as a result of this action. These are all the drive lines connected to the selected

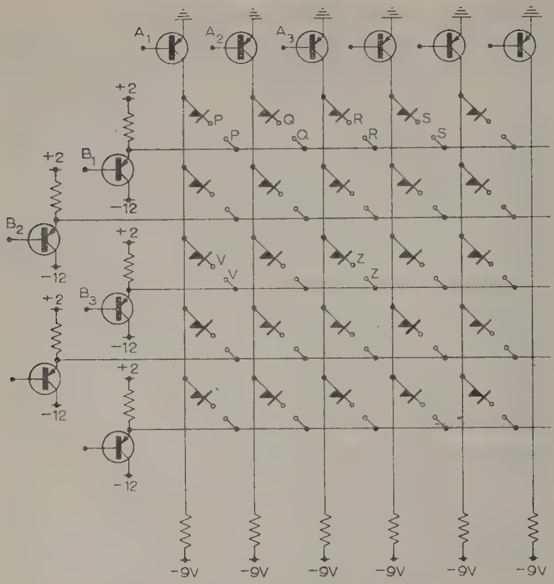


Fig. 15.—Drive-wire selection matrix.

row wire of the matrix, e.g. VV, and these fall from +1 to −9 volts. None of the other drive wires change in potential. The drive lines connected to the selected column wire, such as RR, normally rest near earth potential and experience no change.

The matrix is used to select one out of 256 drive wires and has 16 rows and 16 columns. Because 15 drive lines change potential in addition to the selected line, the matrix system cannot be used in the case of the woven-mesh store without employing transformers. However, the capacitance between drive and read wires is considerably less when the mouldings are used to carry the wires and the present matrix selection system can be used without difficulty. The ratio of magnetic signal amplitude to the capacitive signal is then 100 to 1 in the worst case.

(5) CONCLUSIONS

The storage system described provides a simple and economic technique for the construction of large-capacity, rapid-acces permanent or semi-permanent stores. Two versions of the sto have been constructed and successfully demonstrated. The rea out and cycle times of 100 and 200 millimicrosec are largely dete mined by the bandwidth of the read amplifier. The limiti values are set by the rise time of the current waveform, the pe meability of the ferrite material at very high frequencies and t transmission times for the signals in the store and associat equipment.

The whole or part of the contents of the second version of t store can be changed in under one minute, corresponding to rate of more than 1 500 bits per second.

The primary application of the store is for library programm instructions and permanent data, resulting in an increase overall speed of the computer and making available erasab storage capacity for other purposes. Other applications a envisaged for small special-purpose computers operating wi fixed programmes.

(6) ACKNOWLEDGMENTS

The authors would like to acknowledge the assistance receiv from Messrs. M. Barraclough, K. F. Bowden, J. W. Waterfiel and T. Zombory-Moldovan in the construction of the stor described.

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[The discussion on the above paper will be found on page 605.]

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A PARALLEL ARITHMETIC UNIT USING A SATURATED-TRANSISTOR FAST-CARRY CIRCUIT

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SUMMARY

The paper describes a transistor switch technique which is of particular importance in applications where a large number of switches have to be connected in series and where the propagation time of information through these switches has to be a minimum. It is thus of importance in parallel addition, and its use in this connection has been successfully demonstrated, yielding an addition time over 24 digits of 200 millimicrosec. The technique is reasonably economical, and the paper also shows how it can be used in conjunction with more conventional logical circuits to provide a simple arithmetic unit.

(1) INTRODUCTION

In any programme concerned with solving a problem on a digital computer the computation involved can be divided into three distinct categories. The first is concerned with the main calculation. In a modern universal computer the arithmetic unit provided for such calculations would be expected to operate directly with numbers of the order of 40 to 60 binary digits, this length of number conveniently being termed a word. The main orders would include operations such as shift left or right n positions, various logical operations and also addition, subtraction, multiplication and division of both the fixed and floating-point type.

The second category is concerned with computation on the address part of an instruction which often takes place as that instruction is being obeyed. The number of address digits is defined by the size of the high-speed store, and the capacity of this storage is limited by economic considerations. Thus, in practice, the address is not likely to exceed 20 binary digits, as this number is capable of specifying over one million words. In typical instances the address is confined to a half or quarter word, and therefore arithmetic facilities are essential only over a similar fraction of the word.

The final category of computation defines those small but numerous calculations concerned with the organization of the programme, often termed 'red tape'. This type of calculation, too, can be performed adequately on fractional word lengths.

A single arithmetic unit could deal with all three categories of computation, but during the longer arithmetic orders, e.g. multiplication and division, the high-speed store is capable of supplying information at a rate in excess of that demanded by this unit. Therefore, where speed of computer operation is important, a second arithmetic unit could be provided and its operation overlapped with that of the first unit. If operation of this second unit is restricted to use on the last two categories of computation, the engineering design of the unit can be considerably simplified. For example, no multiplication, division or floating-point facilities are necessary and the numbers involved need only be the appropriate fractions of a word. This simpler form of arithmetic unit forms the basis of the work to be described. It is designed to operate in a parallel manner in order to achieve a maximum speed of operation.

In any parallel arithmetic unit the factor which would appear to limit the speed of operation is the carry signal which has to propagate serially from the least to the most significant channel in the worst case. Numerous techniques have been used in order to minimize the need for carry propagation or to reduce the carry propagation time. Some techniques are enumerated, together with appropriate references:

(*a*) Provide a carry indication at each parallel channel by virtue of gating all the input signals from less significant digit positions (Reference 1).

(*b*) Store the partial answer and a carry in the arithmetic unit and only allow assimilation when the contents of the arithmetic unit have to be returned to the store. This is a particularly useful technique where long sequences of additions occur, as, for example, in multiplication (References 2 and 3).

(*c*) Test for the cessation of carry, since the average distance which a carry propagates is approximately $\log_2 n$, where n is the number length. The time for addition is then variable and is a function of the actual number of channels over which carry is propagated (Reference 4).

(*d*) Split the numbers into groups and test the configuration of digits within the group to determine whether the output carry from a previous group either enters or skips that group (References 5, 6 and 7).

The technique of carry propagation described in this paper reduces the time for carry propagation by such an extent that none of the aforementioned techniques are necessary. The logical simplicity and economic advantages of this new system are also significant points in its favour. A recent short paper[8] gave an intimation of the technique used. That discussion is now further expanded and more practical details are presented.

(2) LOGICAL DESIGN OF THE ADDER

The design of stage k of an n-stage parallel adder using mechanical switches is shown in Fig. 1. The kth digits of the numbers x and y and the carry signal from the $(k-1)$th stage are added to produce the sum A_k and the carry information C_k and $\bar{C}_k$. The switches are controlled by an appropriate logical function generated from the x and y digits particular to that stage. A signal originates when the x_k and y_k digits are equivalent; two '1's generate a carry and two '0's a not-carry signal. These signals are propagated along two separate paths termed the 'lower carry path' and 'upper carry path', respectively. Signals similar to these but arising in less significant stages can be transmitted through stage k when x_k and y_k are not equivalent. In any stage the sum output is formed by routing the input carry and not-carry to a common output via switches which operate in a mutually exclusive manner.

The control of these switches is independent of the nature of the carry signals, and thus all switches can be operated simultaneously in a time referred to as the 'set' time t_s. Subsequently the maximum delay in determining the value of the most significant answer digit A_{n-1} is that appropriate to the propagation

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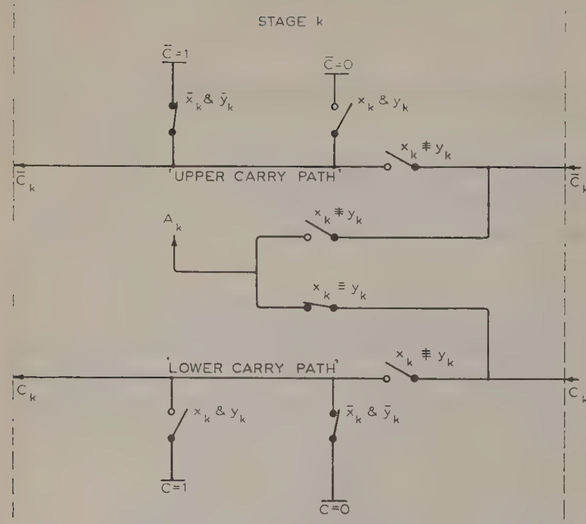

Fig. 1.—Logical diagram of parallel adder using a switch path.

of a voltage step through $n - 2$ contacts in series plus their interconnecting wire. Since the switches are of a mechanical type, the speed of propagation approaches that of light, and the actual delay is calculable from the physical length of the path. The technique thus makes the carry propagation delay small compared with the 'set' time of the switches, which therefore becomes predominant in the determination of the time for an addition. This 'set' time of the switches can be reduced in practice by the use of junction transistors which are used in their no-current and 'saturated' states to simulate the open and closed positions of the mechanical switch. An important property of a saturated transistor is that a voltage step can be passed from emitter to collector with negligible delay, provided that no change in the saturation condition occurs. This feature is independent of the cut-off frequency f of the transistor, but it is important to have a high value of f in order to switch the transistor in and out of saturation at a high rate. The use of transistors has thus maintained the carry propagation time at a small value but decreased the 'set' time and hence the addition time very significantly.

In practice the original design has been modified to that shown in Fig. 2. The upper carry path has been eliminated for economic reasons, and the sum switches have been isolated from the carry path by using emitter-followers in order to minimize loading on that path. The sum switches, too, have been changed to the more conventional diode gates, since that form of circuit can be accommodated more conveniently in the part of the arithmetic unit dealing with the sum output, and maximum speed of propagation is not as important in these positions. Further information on the transistor switch is now provided.

(3) THE TRANSISTOR SWITCH

The simple equivalent circuit of a *p–n–p* transistor is shown in Fig. 3. The diodes indicate the rectifying nature of a *p–n* junction, and the current generators F and R provide currents to simulate transistor action when the emitter-base and collector-base junctions, respectively, are forward biased. Most transistors are not symmetrical and typically $\alpha_R \ll \alpha_F$. The condensers shown vary according to the operating condition of the junction. When reverse biased the capacitance is referred to as

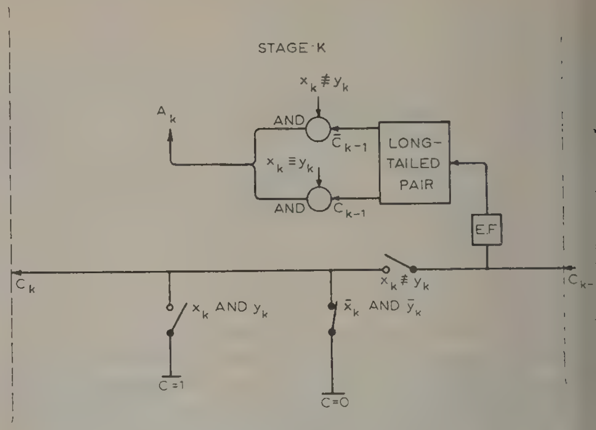

Fig. 2.—Simplified diagram of carry path and sum logic.

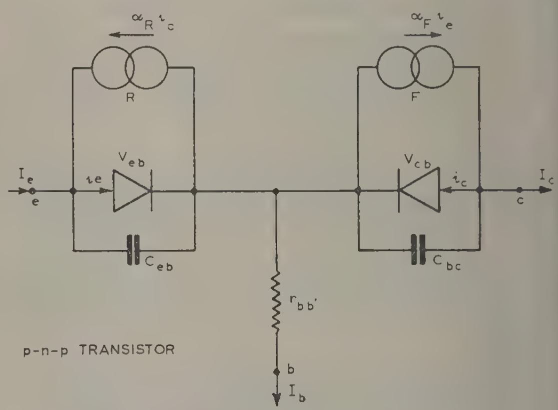

Fig. 3.—Simple transistor equivalent circuit.

a barrier capacitance and its value is approximately inversel proportional to the square root of the voltage across it. I the forward-biased condition a diffusion capacitance exits whic is proportional to the forward current, and except in v.h.f transistors ($f = 1$ Gc/s) this diffusion capacitance is muc greater than the barrier capacitance. Two forms of transisto switch exist, and these are identified by the bias conditions o the two junctions.

In the conventional switch the collector-base junction i reverse biased, and the collector current I_c is dependent upo the amount of forward bias applied to the emitter-base junctior If this bias is changed the number of carriers crossing the junctio increase, but before they reach the collector some time mus elapse for these carriers to diffuse across the base region. Th transistor has to change its condition in order to pass informa tion, and the switching time t_g is a function of f, the transisto cut-off frequency. If n of these switches are connected in serie the time to propagate information from one end to the other i proportional to nt_g, i.e. n/f.

The second form of switch is similar to a mechanical switc in that its contacts can be closed to complete a circuit inde pendently of the passage of information. The time to pas information through n such contacts in series is then equal t $t_s + nt_p$. The term t_s is the 'set' time of the slowest switch and, as can be seen later, is a function of f, although in practic t_s will always be greater than t_g. The other factor t_p is th

·opagation time of information through one contact, and if is is similar to a mechanical contact the velocity of propagation ill approach the velocity of light. It is possible, in practice, make nt_p and t_s small, so that $t_s + nt_p$ is less than nt_g, and this way the carry process can be accelerated.

The 'off' condition of this switch occurs when both junctions e reverse biased, and the only current which then flows will due to leakage. The 'on' condition corresponds to both nctions forward biased, a state which occurs when $\beta I_B > I_c$. is the external collector current which can be limited by the cternal circuit, and β is the usual current gain defined as $/(1-\alpha)$. Under this condition some current i_c from the enerator providing $\alpha_F i_e$ flows through the diode V_{cb}. Since $I_B > I_c$ many more carriers cross the emitter-base junction ian are collected by the collector, and these are stored in the ase region (hole storage). The transistor in this state is said be saturated.

Provided that the state of saturation is maintained, a very low npedance exists between emitter and collector, and a step of irrent or voltage can be transmitted almost instantaneously om emitter to collector. The voltage drop across the transistor nitter to collector corresponds to the difference in voltage drop cross V_{eb} and V_{bc}, and this can be very small as confirmed by ie experimental results given in Appendix 12.1. This low oltage drop permits many such switches to be connected in eries.

Switching a transistor out of saturation takes longer than witching off a similar current in a normally operated transistor. his is because a large diffusion capacitance exists across both nctions instead of only the emitter-base junction. Any icrease in the charge supplied to the base terminal will nprove the switching time.

The diffusion capacitance is proportional to $1/f$, and thus a igh-frequency transistor is required to achieve the fastest 'set' me. Available transistor types are as follows:

Alloy: f up to 20 Mc/s.
Surface barrier: f up to 40 Mc/s.
Drift: f up to 100 Mc/s.
Micro alloy diffused: $f = 250$ Mc/s.
Mesa: $f = 600$ Mc/s.

Most of the drift and mesa types have the characteristic that ieir high-frequency performance is only obtained with a reverse ollector-base voltage of about 3 volts, and thus they do not perate at high speed in and out of the saturated state. Switch-ig tests carried out on the other types of transistor are discussed Appendix 12.2.

Switching time is also affected by the connection of the ansistor, i.e. whether it is switched in the forward or reverse irection. The latter case corresponds to use of the collector as n emitter. Any transistor with a built-in 'drift' field will witch slowly in the reverse mode, and hence in an application here this mode is used, a similar performance might be obtained a more economical manner by using a slower but more sym-etrical transistor. Thus, for example, the surface-barrier ansistor 2B240 and the micro-alloy diffused transistor 2N501 re almost equivalent if the sum of their forward and reverse witching times is taken as the figure of merit.

The switching conditions in the carry path are now discussed.

(4) AN EXPERIMENTAL CARRY PATH

One stage of the practical carry path is shown in Fig. 4. This iagram closely resembles Fig. 2, but now the mechanical switch ontacts have been replaced by transistor circuits. The tran-stor switches are controlled by function waveforms applied irough a diode-resistor circuit to the base terminal. This rrangement ensures that the transistor switches on with a relatively constant current which is independent of small voltage movements of emitter, base and collector. The reverse current available to switch the transistor off is a function of the source impedance driving the function waveform. These function waveforms operate transistors T, T′ and T″ in a mutually exclusive manner so that only one of the transistors is defining the voltage level of the carry path. Under transient and fault conditions, however, two transistors could be switched on together, for example T′ and T″ conducting simultaneously would try to connect the −3 and −4·5-volt lines together. To prevent excessive power dissipation in the transistors in this instance the −3-volt line is limited to a maximum current of 15 mA.

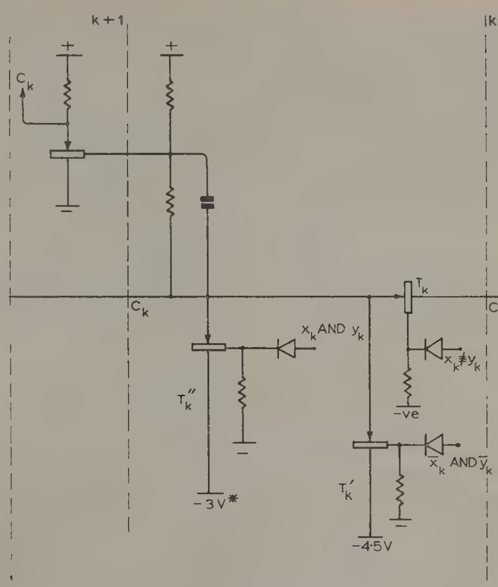

Fig. 4.—Practical carry path.

When the adder is inactive the T′ transistors are all conducting to hold the carry path at −4·5 volts, the no-carry voltage level. The emitter and base currents are almost equal to ensure that a minimum current is demanded from neighbouring stages by the transistors. To establish a carry, for example, in stage k, the T'_k transistor is turned off and the T''_k transistor is turned on, thus changing the voltage of the carry path to −3 volts, the 'carry = 1' level. The amplitude of the carry signal is 1·5 volts, and this has been made small to minimize the amplitude of the function waveforms which must completely envelop the carry signal in order to switch the transistors correctly. The function wave-forms swing between −5 and −2 volts, respectively.

The function waveforms have a finite rise time, and when considering a change in the setting of switch transistors T, T′ and T″, significant delays can be caused by the function waveform moving to a suitable level for operating the transistor. Further variation occurs because the transistors T and T″ can both be switched using the collector as an emitter. When the emitter is more positive than the collector at the time of switching, the speed of switch is governed by f_F, the α cut-off frequency of the current generator F. However, if the collector is more positive, the cut-off frequency f_R of the current generator R will be the main factor. If $f_F \gg f_R$ it is desirable to use some form of reset technique so that forward switching is always ensured. Alternatively a symmetrical transistor should be used in this position.

When a number of these stages are operating, the stage initiating the carry has to be capable of driving the other stages plus their associated capacitance. In practice, the speed of switching the carry-generating transistors can be improved by inserting an emitter-follower after each group of six stages. An emitter-follower stage has a potential difference between emitter and base of the order of 0·3 volt in the type of transistor used, and this causes a shift in the level of the signal. Since the carry signal is only 1·5 volts in amplitude, a shift of 0·3 volt every six stages is undesirable. A diode of the gold-bonded type is therefore used to back off this voltage drop by lowering the output signal level from the emitter-follower. Though, in practice, this cannot be done exactly it has been found dependable in reducing the drop.

Twenty stages of this carry path have been constructed and operated satisfactorily using SB240 transistors in positions T, T′ and T″. A carry could be transferred from the least to most significant in about 80 millimicrosec. The delay was found to be made up of two components. The first depends on the rise time of the function waveforms and the setting of the transistor switches, which, of course, is not a cumulative effect. The second component was the delay through the closed switches principally due to the physical length and form of the path. This length was approximately 5 ft and resulted in a delay of about 20 millimicrosec. Replacement of the T transistors by short loops of wire showed that the transistors in this configuration contributed some 10 millimicrosec of delay.

This carry path was incorporated in an experimental adder for further test.

(5) EXPERIMENTAL ADDER

The logical arrangement of this adder is shown in Fig. 5. Numbers from the x and y registers can be added together and

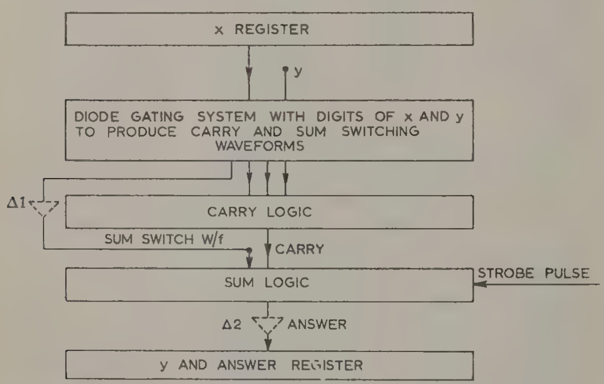

Fig. 5.—Schematic of adder system for operation $x + y$ to y.

the result returned to the y register. The y register thus contains a ‘running’ total.

The x and y register outputs drive diode gates generating the sum and carry function waveforms that are defined in Fig. 2. The gate outputs were emitter-followed in order to drive the carry and sum logic. There are two routes through the adder which can affect the output from the sum logic. The short route which circumvents the carry path is due to the sum switch waveforms, and this is assumed to have a delay t. The route through the carry path has a variable delay θ_{min} to θ_{max} which depends on the carry signal arising in either the most or the least significant stage. The difference $(\theta_{max} - \theta_{min})$ is the propagation time of the carry from the least to the mo significant position.

The output from the sum logic is gated to the y register by narrow strobe pulse of duration τ, which must be less than t minimum delay through the adder. This must occur in ord that a change in the output of a y register cannot feed back time to be included within the strobe duration. The minimu time between strobe pulses, i.e. the addition cycle, must greater than the maximum delay through the adder.

In practice τ is determined by the set time of the answer fli flops, and t, θ_{min} and θ_{max} are also known. Thus if $\tau > t$ θ_{min} and $t < \theta_{min}$, an extra delay Δ_1 must be included as shown Fig. 5, so that $(\Delta_1 + t) = \theta_{min}$. Since $\tau > \theta_{min}$ a further del Δ_2 will have to be inserted in the common path so th $\tau \leqslant \theta_{min} + \Delta_2$. The time between additions is now $\theta_{max} + \Delta$

In the experimental adder τ was 100 millimicrosec. Δ_2 w also 100 millimicrosec, and an addition cycle of 200 millimicros was achieved. Since the operation of this adder, faster flip-fl circuits have been designed, and the length of the carry pa has been reduced by careful layout on the printed circuit board These factors should contribute some improvement in speed, b in any arithmetic unit there is likely to be more complexity th in the system described, and this will detract from the improv ment. An arithmetic unit is described in the next Section.

(6) SIMPLE ARITHMETIC UNIT

The simple arithmetic unit carries out arithmetic operatio and address modification on half words of 24 binary digits. is termed the B accumulator in accordance with the usual nome clature for address modifiers. The unit differs from the ma accumulator in that a large number of answer registers a provided. In the development of machines the number of registers has increased from 2 to 4 to 8 and 16 respectively, t larger number of B registers being particularly useful in aut code procedures. It is thought that B registers will also be demand for machines using extensive time-sharing techniqu to peripheral equipment, and for this reason 128 such registe have been provided in this unit. To be economic these registe are constructed in the form of a 128-word core storage syste with a cycle time[9] of approximately 0·5 microsec.

A list of the instructions associated with this unit is giv in Appendix 12.4. It contains the usual logical and arithme facilities plus two shift orders which facilitate manipulation six-digit characters and provide access to single digits in ea character.

A block diagram of the unit is shown in Fig. 6. The tw numbers which enter the unit from the store S and register respectively, can be selected from one of several sources, a these channels are shown gated together into the S_k and information channels. The S_k number is used directly, but t B_k number may be shifted either six places to the left (mo significant) or one place to the right (less significant). Bo these shifts are of a circular nature. The input information and B_k or B_{k-6} or B_{k+1} is then converted to double pha signals which are used in either the logical operations unit the adder subtractor. It would be possible to gate the adder/su tractor to provide logical operations, but to avoid slowing down, a completely separate logical circuit has been provide Control signals enter both the adder/subtractor and logical u to define the required operation. Subsequently the outputs each unit are gated together to provide the final output. T may be sent either to an S- or a B-type register, and to achie fast operation both phases of output are provided. This tec nique eliminates the need for a reset to the S or B registers.

The realization of this arithmetic unit in practical terms

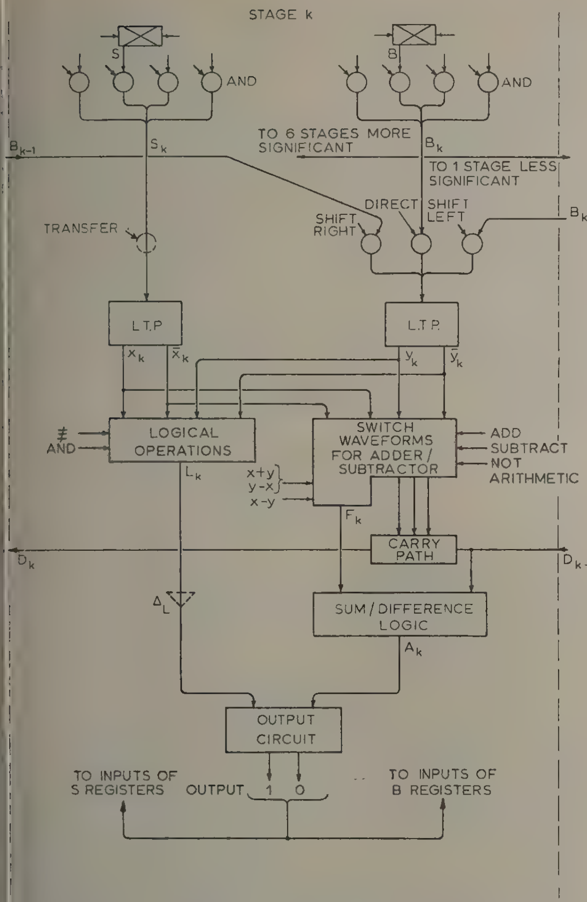

Fig. 6.—Block schematic of one stage of simple arithmetic unit.

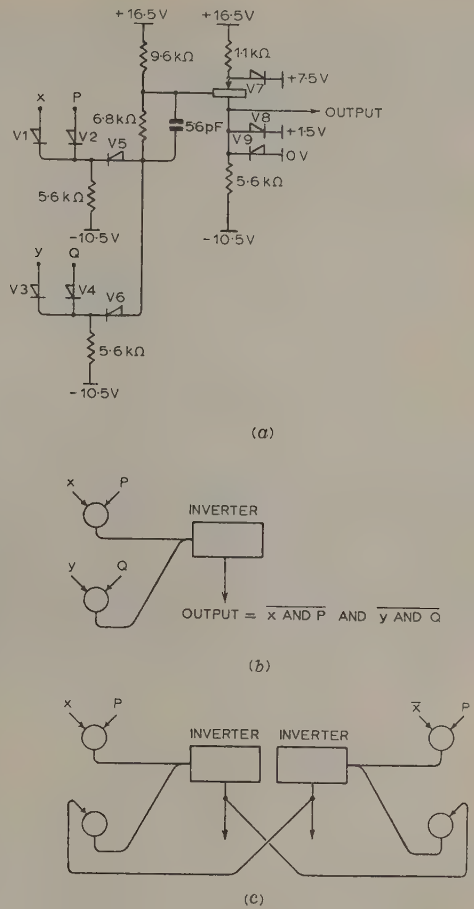

Fig. 7.—Standard circuit technique.

(*a*) Practical circuit.
(*b*) Logical representation.
(*c*) A standard register.

ependent upon the circuit techniques used for the gating nd storage requirements. These techniques are now briefly scussed.

(6.1) Circuit Design

In the major part of the machine standard logical units are sed to provide gating, amplification and restandardization of gnal voltage levels. The carry circuit, however, produces an itput of non-standard form since this carry signal must be mpletely enveloped by the function waveforms. Hence the m logic has been designed to accept non-standard input levels id provide standard output signals.

1.1) Basic Logical Unit.

Initial work on the experimental adder showed that it was sirable to define clearly the voltage levels of the signal and ting waveforms. The basic logical unit shown in Fig. 7(*a*) s been designed to provide this feature. It consists of a ansistor amplifier with a defined emitter current of 8 mA. is current flows into a catching diode V_7 when the transistor cut off and into the collector when the transistor is switched . The collector voltage levels are defined by two diodes V_8 d V_9 which are attached to the clipping levels of 0 and 1·5 volts. These diodes are of the point-contact low-hole- rage type and allow the collector voltage to swing between 0·5 and +2 volts. The current available at these levels depends on the collector load, but is a maximum of 7 mA, which can be provided as $\pm 3\cdot 5$ mA.

The input circuit consists of two or more diode 'and' gates which are followed by an 'or' gate. The information signals *x* and *y* are typically at collector levels, and the necessary change in d.c. level to drive the base of the transistor is achieved by the use of a resistance network. The condenser is calculated to supply sufficient charge to the transistor base to switch the emitter current completely on or off.

The large voltage drop from emitter to collector in excess of 5 volts is maintained to provide good high-frequency characteristics in the transistor. It means, however, that the resistance networks to the base must be defined with 1% resistors. A very stable metal-oxide resistor type is used. In practice, the typical delay through a stage is of the order of 15 millimicrosec, and the rise and fall times are about 20 millimicrosec. These measurements may be found to be in error when a better oscillograph is available. The logical unit can in theory drive three diode 'and' gates, but, in practice, to achieve the maximum speed the load is restricted to a maximum of two gates.

This may at first sight appear restrictive, since, in a computer, an output may be connected to many inputs. However, at any given time only one and sometimes two inputs are activated, which means that the d.c. loading of a multiple connection can

be reduced by making the control waveforms P and Q to the 'and' gates be at a level where they take all the current. These larger control waveforms are distributed by special driver circuits. In this way all the non-operative connections only increase the capacitance applied to the output of the logical unit.

The logical representation of this unit is shown in Fig. 7(*b*), and in Fig. 7(*c*) the necessary arrangement of two such units to make a flip-flop is demonstrated. This flip-flop is set or reset according to the phase of x and $\bar{x}$ when the control waveform P goes negative. A further modification to the unit, replacing V_7 by a transistor with a common-emitter connection, base connection to $+7\cdot5$ volts and identical collector circuit, produces a unit providing both phases of output information and is termed a 'long-tailed pair'. These units and the carry path already discussed are used in the design of the arithmetic unit.

(6.2) Detailed Design of the Arithmetic Unit

The input gates of the arithmetic unit are shown in Fig. 6. The signals (S_k, B_k) are routed through and gates to long-tailed-pair circuits which provide both phases of information signals x_k, $\bar{x}_k$, y_k and $\bar{y}_k$. In the quiescent state with all gates closed by control signals the x_k and y_k outputs will be at the more positive voltage, thus representing '0'. The outputs from these long-tailed-pair circuits are used to perform addition, subtraction and logical operations, and the design of these computing circuits is conditioned by the fact that one output must not drive more than two gates at any time. The way this has been achieved is shown in Fig. 8.

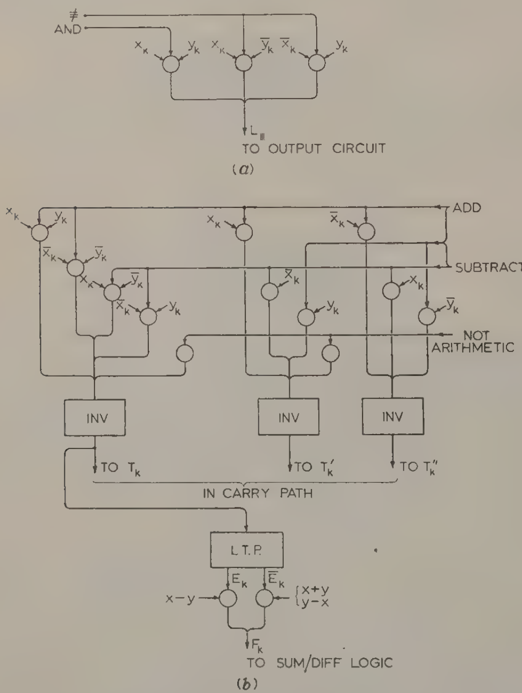

Fig. 8.—Detailed logic design of arithmetic unit.

(*a*) Logical operations.
(*b*) Switch waveforms for adder subtractor.

The logical functions 'and' and 'not equivalent' are produced directly by the circuit of Fig. 8(*a*), and the operation 'or' may also be carried out by simply applying the 'and' and '≢' control waveforms simultaneously. When these operations are being performed the arithmetic gates are isolated from the information signals (x, y, $\bar{x}$, $\bar{y}$) by control waveforms, and similarly durin[g] addition or subtraction the 'and' and '≢' control waveforms ar[e] both more positive than these signals. Generation of th[e] switching waveforms applied to the transistors T_k, T'_k and T[''] in the carry path is illustrated in Fig. 8(*b*). The logical desig[n] is more complicated than that previously discussed since th[e] carry path is used for addition and subtraction. When logica[l] operations are being performed the control waveforms 'add' and 'subtract' are positive to isolate the d.c. loading of th[e] adder/subtractor gates from the long-tailed-pair circuits pro[-]viding the information signals. Under these conditions selectio[n] of T_k, T'_k and T''_k in the carry path would no longer be mutuall[y] exclusive. To prevent this, the carry path is established at [a] standard state by the control waveform 'not arithmetic'. Thu[s] three modes of switching occur in the carry path and the wave[-]forms applied to the transistors T_k, T'_k and T''_k under eac[h] circumstance are as shown in Table 1.

Table 1

FUNCTIONS APPLIED TO CLOSE TRANSISTOR SWITCHES T_k, T'_k AN[D] T''_k IN THE CARRY PATH

Operation	T_k	T'_k	T''_k
Addition ..	$x_k \not\equiv y_k$	$\bar{x}_k$ and $\bar{y}_k$	x_k and y_k
Subtraction ..	$x_k \equiv y_k$	x_k and $\bar{y}_k$	$\bar{x}_k$ and y_k
Not arithmetic	Open	Open	Closed

Three arithmetic operations $(x + y)$, $(x - y)$ and $(y - x)$ ar[e] to be carried out. In switching the carry path, different signa[ls] only arise from the statements 'add' or 'subtract'. To achiev[e] this, the levels of the carry signals must be interpreted in [a] different sense according to the type of subtraction performe[d]. A negative level indicates 'carry or borrow = 1' for $(x + y)$ an[d] $(x - y)$, respectively, and a positive voltage level a 'borrow = 1' for the $(y - x)$ operation. This statement is summarized i[n] Table 2.

Table 2

FUNCTIONS USED IN SUM/DIFFERENCE LOGIC

Operation	Carry signal D_{k-1}	E_k	F_k
$x + y$	C_{k-1}	$x_k \not\equiv y_k$	$x_k \equiv y_k$
$x - y$	W_{k-1}	$x_k \equiv y_k$	$x_k \equiv y_k$
$y - x$	W_{k-1}	$x_k \equiv y_k$	$x_k \not\equiv y_k$

The signal which, when combined with the carry, provide[s] the sum or difference output is generated from the switchin[g] waveform applied to T_k. Two phases E_k and $\bar{E}_k$ (see Fig. [8]) are generated by a long-tailed pair, and these are gated wit[h] signals specifying the type of operation to provide a signal F_k. The form of signals E_k and F_k for the different operations [is] given in Table 2.

The design of the circuit which combines waveform F_k an[d] the output of the carry path, D_{k-1}, to provide the sum [or] difference is shown in Fig. 9. This Figure also shows the outpu[t] circuit which combines the logic and adder/subtractor output[s]. During arithmetic operations the signal L_k is at a positiv[e] voltage level and thus does not affect the output. The signa[l]

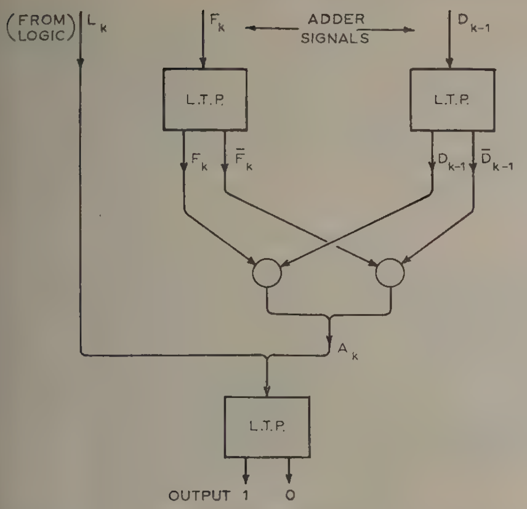

Fig. 9.—Design of sum/difference logic and output circuit.

$_{-1}$ and F_k are supplied in a double phase form to the gates ich provide the arithmetic sum or difference, A_k.

This function can be expressed by the following Boolean uation:

$$A_k = (F_k \text{ and } D_{k-1}) \text{ or } (\bar{F}_k \text{ and } \bar{D}_{k-1})$$

hen this equation is compared with Table 2 it is evident that equations describing the sum and difference can both be eyed.

Sum $= (x_k \equiv y_k)$ and c_{k-1} or $(x_k \not\equiv y_k)$ and $\bar{c}_{k-1}$.
Difference $= (x_k \equiv y_k)$ and w_{k-1} or $(x_k \not\equiv y_k)$ and $\bar{w}_{k-1}$.

If the adder/subtractor is inoperative F_k is always positive and $_{-1}$ is negative, thus ensuring that A_k indicates zero. The tput of the adder/subtractor is combined with that of the ical circuits in an 'or' gate which feeds a long-tailed-pair cuit producing a double-phase output signal.

.1) Practical Details of the Carry Path.

The reversal in significance of the voltage levels of the carry th to accommodate both types of subtraction is more eco- mical and provides slightly faster waveforms to the switches the carry path than the alternative methods. It results in ne complication to the least significant stage of the carry h, but in all other stages the two forms of subtraction need t be identified. In stage '0' of any adder/subtractor, carry or rrow signals are only generated, and thus, in this instance, y two transistor T_0'' and T_0' are essential. However, to nplify the switching waveforms it is convenient to have the angement shown in Fig. 10. In stage '0' the transistors are ntrolled by waveforms identical with those in other stages and ed in Table 1. Thus during addition and subtraction the nsistor T_0 is switched by $x_0 \not\equiv y_0$ and $x_0 \equiv y_0$, respectively, l these conditions should transmit a 'carry = 0' or 'borrow = signal through stage '0'. Since the level of the carry path be interpreted in different ways, dependent upon the mode subtraction, two transistors T_{-1}'' and T_{-1}' attached to different els have to be used. These are switched, as shown by the propriate function waveforms. In this way it is also possible perform the operation $x + y + 1$.

The output levels from the collector of a standard inverter are proximately $+2$ and $-0{\cdot}5$ volts, respectively. The upper el is inadequate to switch the transistors of the carry path,

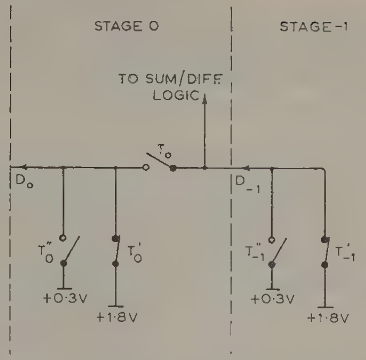

Fig. 10.—First stages of the carry path.
T_{-1}'' closed for operations $(y - x)$ or $(x + y + 1)$ or (not arithmetic).
T_{-1}' closed for operations $(x + y)$ or $(x - y)$.

owing to voltage variations along the path, and is increased to $+2{\cdot}5$ volts at the outputs of the inverters driving the switch transistors. The carry signal levels, nominally $+1{\cdot}8$ to $+0{\cdot}3$ volts, lie within those of the switching waveforms. This non-standard amplitude and level means that the carry signal cannot be applied directly to a basic logical unit. The sum/difference logic shown in Fig. 9 is therefore performed by non-standard circuit techniques described in Section 12.3.2.

(6.2.2) Inherent Delays in the Arithmetic Unit.

The delays in the arithmetic unit are of two types, and these occur in the basic logical inverter circuit and the carry path, respectively. The former delay in circuits with a minimum-capacitance load varies between 12 and 20 millimicrosec. In practice, with capacitance loading these figures can be increased to 15 and 25 millimicrosec. The carry-path delay consists of the set time of the slowest switch, approximately 80 millimicrosec, and the propagation delay of 20 millimicrosec. This former time is greater than the value quoted in Section 12.2 owing to the increased capacitance both on the carry path and the stages providing the switch waveforms to the path. These capacitances arise because of

(*a*) The presence of a third switch transistor and an emitter-follower at each stage.
(*b*) The presence of many stages of carry.
(*c*) Wiring between boards.

In the arithmetic unit two delays are of interest, one through the path which performs the logical operations and the other through the adder subtractor. In the former instance delay from input to output arises due to four basic stages through which information passes, and this delay is a minimum of 48 millimicrosec. For the latter case the minimum delay from input to output can occur via the sum switch waveforms which by-pass the carry path as explained in Fig. 5; this amounts to 72 millimicrosec, i.e. six invertor stages. The strobe-width gating information into flip-flops is defined by the minimum delay through the arithmetic unit, and to make this consistent for both logic and addition or subtraction, a delay ΔL is included in the logical-operations channel as shown in Fig. 6. The minimum time between successive strobe pulses is defined by seven times the maximum delay in a basic circuit plus the carry-path delay. This amounts to 275 millimicrosec.

In the main accumulator of the machine there will be a 40-stage adder. Assuming that 40 stages of carry path can be connected in series, this increase in the number of stages only contributes an extra delay of 14 millimicrosec. However, since the accumulator does not perform the logical operation and also the number

of inputs to be gated is less, two basic circuits can be removed from this path and the maximum accumulator addition time is expected to be of the order of 250 millimicrosec. In the main accumulator multiplication and division have to be performed and these operations are now discussed briefly.

(7) MULTIPLICATION AND DIVISION

The time for multiplication of two numbers D (multiplicand) and R (multiplier) is dependent upon the number of sub-products which have to be dealt with, and this is clearly a function of number length. In the simplest case if D and R are 40 digits the number of sub-products is also 40. This number of sub-products can be reduced by using the following method. From the register storing D, the numbers D, $2D$ and $4D$ are readily available. If a further register is provided a single addition of D and $2D$ provides the number $3D$. Thus the multiplier number R can be examined in groups of three digits and the appropriate sub-product 0, D, $2D$, $3D$ or $4D$ added or subtracted from the running total of sub-products. In this way only 39/3 sub-products need be provided and the multiplication time is thus $13 + 1$ addition times, i.e. approximately 3·5 microsec. The built-in shift in the accumulator will then be over three positions. If the shift time is less than an addition time which is doubtful at this stage, a further time reduction could be achieved by merely shifting for groups of three '0's or three '1's.

Division also requires a number of repeated subtraction or addition operations. Since division is a process of trial and error, a maximum of 40 such operations could be required, although a small amount of test equipment permits the average number of operations to be halved.

(8) ANOTHER APPLICATION OF THE SWITCH—STANDARDIZATION OF NUMBERS

In floating-point machines numbers typically consist of two parts, a fraction X and exponent Y. In floating binary the range of numbers which can be represented is given as $X2^Y$, the fraction X being specified to a certain accuracy within the range of positions denoted by 2^Y. The fraction X has a 'standard form' representing a range of numbers which the arithmetic unit accommodates. When a number exceeds the range, for example by a carry into a more significant place, the accumulator is shifted to bring the number within range again and the exponent is appropriately modified. This process is termed standardization. In a very restricted range of numbers, for example, a number is said to be standard if it exhibits a change-over in the value of digits occupying the most significant and the next-to-the-most-significant digit positions.[10]

Standardization can be carried out by a shift and test process, the direction of shift either to least or most significant being determined by another test. When a small number has to be standardized the successive shift and test procedure can take a long time. If the accumulator is provided with multiple shift paths, it will be faster though more expensive to detect the most significant change-over of digit values and then carry out the appropriate shift and exponent correction.

Circuits to detect the most significant change-over in this way are illustrated in Fig. 11. The switch path shown indicates a voltage M only on the output point in the most-significant-digit position at which $x = 1$. All other positions will have the voltage 0 on the output point. Consider how the switch path shown reacts to the following positive input number:

$$0{\cdot}01\delta\delta\delta\ldots$$

In the first two digit positions the path will show voltage M, but the output will show voltage 0 since the switches X will be

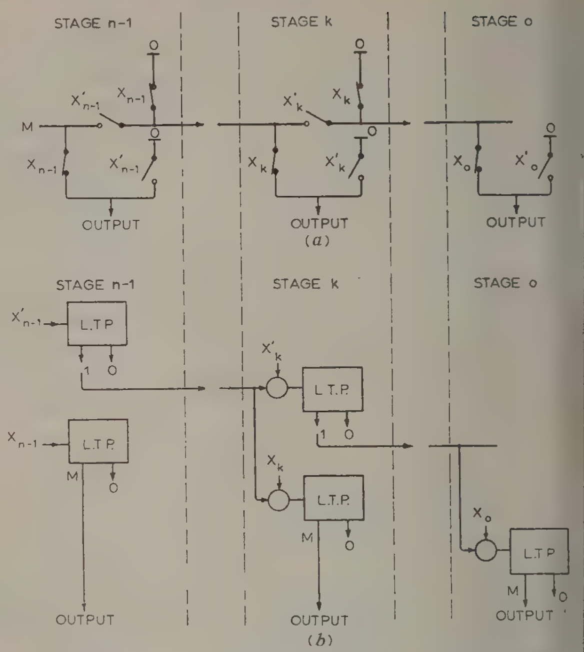

Fig. 11.—Most significant change-over detectors.

(a) Switch path.
(b) Standard logical circuits.
$X = x$ for positive number or $\bar{x}$ for negative number.
$X' = \bar{x}$ for positive number or x for negative number.
Output $= M$ at stage following most significant change-over.

open and X′ will be closed. The third digit position will intr duce voltage 0 to the path, but its output will be voltage M sin switch X′ is open and X closed. Further digit positions ca now only ever have voltage 0 on the path and can therefo only ever give an output voltage 0 irrespective of the order which the switches open and close. A negative number w have a change over from '1' to '0' instead of from '0' to '1 It is therefore necessary to change the significance of the signa controlling the switches. A method using standard logic circuits is shown in Fig. 11(b). The signal M may have to pa through all the 'and' gates corresponding to all the digit position The delay may be as much as 800 millimicrosec in a 40-dig number. This compares very unfavourably with the delay of le than 150 millimicrosec expected along the switch path and emph sizes the superiority of the switch path technique.

(9) CONCLUSIONS

A useful transistor switch technique has been demonstrat which is primarily of importance in reducing carry propagati time in parallel addition. The time taken for a carry to reach t most significant stage is a function of the length of path and t switch time of the transistors in and out of saturation. T former time is of the order of 20 millimicrosec in the final schem discussed, and this might be further improved by special layo schemes for the switch resistors. Wire lengths on the transist emitter and collector connections are also of importance, a these too could be reduced by selection of the right transist type. The latter time is a function of the speed of edge of t controlling waveforms to the switch transistor and also of t switch transistor itself. A maximum switch time including bo these factors is 80 millimicrosec.

At this stage of development the main factor determining t addition time is the speed of the conventional logical un

ection 6.2.2 shows that a time of 175 millimicrosec is due to ese units and only 100 millimicrosec to the carry path. Thus a gnificant improvement in addition time can be achieved by eeding up operation of the logical unit. This suggests the e of transistors such as the 2N1142 which should provide a eed improvement by a factor of at least two. Under these rcumstances an addition time of better than 175 millimicrosec uld be achieved.

(10) ACKNOWLEDGMENTS

The authors wish to thank Ferranti, Ltd., for facilities provided d also all members of the Atlas Computer Research Team, th at Ferranti, Ltd., and Manchester University, for helpful scussion and other contributions to the work described. cknowledgment is particularly made to Mr. E. T. Warburton r his design effort on the logical unit, and to Mr. E. C. Chen r his considerable help with measurements and constructional tails. Mr. R. A. Brooker is also to be thanked for preliminary scussions concerning the logical design of parallel adders.

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(12) APPENDICES

(12.1) D.C. Measurements on Switch Transistors

The circuit of the carry path must be designed so that the tch transistors have a low potential difference between emitter l collector when saturated. Measurements of this potential erence have been made on SB240 and 2N501 transistors.

Two series of measurements are described. In the first the lector is held at a fixed voltage and the base current made al to the emitter current. Ten samples of each type of nsistor are tested and the results are given in Table 3(*a*). The ential difference is measured by a comparison method which is urate to within $\pm 0\cdot 1$ mV. The results show that the SB240

Table 3

Potential Difference between Emitter and Collector of SB240 and 2N501 Transistors

Transistor type	Potential difference between emitter and collector, mV					
	Max.	Av.	Min.	Max.	Av.	Min.
SB240	10·5	8	6·4	3·2	2·8	2·2
2N501	33·3	24	15·2	54	7	0·4
	$I_c = 0$, $I_e = I_b = 1$ mA			$I_e = 0$, $I_c = I_b = 1$ mA		
	(*a*)			(*b*)		

has a lower potential difference than the 2N501. In the second series the emitter is held at a fixed voltage and the base current made equal to the collector current. The same batch of transistors is used and the results are given in Table 3(*b*).

The SB240 transistor has a lower potential difference between emitter and collector when collector current equals base current and the emitter current is zero, than when emitter current equals the base current. This is explained by the results of a third series of experiments in which the transistor is connected with its collector at a fixed voltage and with a constant base current of 1 mA.

The emitter current is varied and the potential difference between base and collector, V_{bc}, and between base and emitter, V_{be}, are measured by a comparison method. The results for the SB240 and 2N501 transistors are plotted in Fig. 12. Each

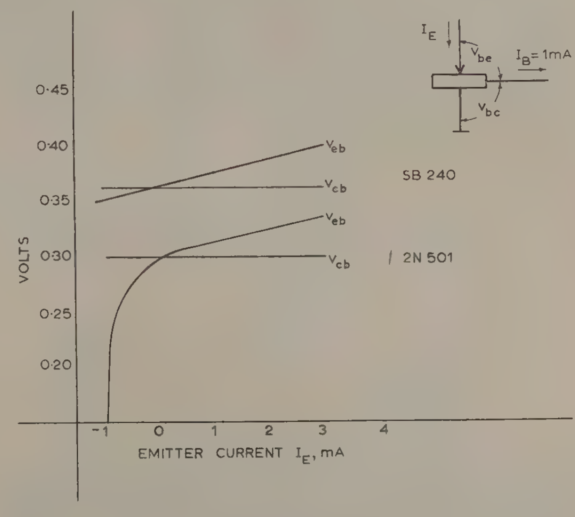

Fig. 12.—D.C. measurements on SB240 and 2N501 transistors in saturation.

transistor has a value of V_{cb} which shows little variation in the range of currents indicated. In each case V_{eb} is greater than V_{cb} for positive values of emitter current, and yet less than V_{cb} for negative emitter current. Thus the difference between V_{eb} and V_{cb}, which corresponds to the potential difference between emitter and collector, is zero in the region of zero emitter current.

The 2N501 transistor shows an abrupt change in V_{cb} as the emitter current becomes negative. This causes the potential difference between emitter and collector to be very sensitive to current in the region of zero emitter current and accounts for the wide range of results recorded in Table 3(*b*).

This Table shows that the transistor with the lowest average voltage drop is the SB240 connected so that the collector current equals the base current.

(12.2) **Switch Measurements on Transistors**

A circuit which simulates the condition of the switch path is used to test the switching properties of transistors and is shown in Fig. 13. The transistors under test are placed in the positions

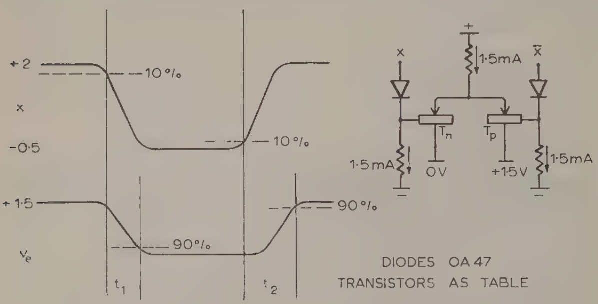

Fig. 13.—Test circuit. Switch properties of transistors.

T_n and T_p. The switching waveforms x and $\bar{x}$ are derived from a standard long-tailed-pair unit which thus provides output levels of +2 and −0·5 volts. The timing measurements are carried out with reference to the x waveform, and the method of measurement of times t_1 and t_2 is shown in Fig. 13.

A batch of 10 transistors was selected at random, and whilst one was placed in position T_n the nine others were placed successively in position T_p when the measurements were taken. The original T_n transistor was then inserted in position T_p and the measurements were repeated inserting the nine transistors into the T_n position in turn. The results shown in Table 4 are the mean values obtained as a result of these 18 measurements on each of several transistor types.

Table 4

Switch Times of Various Transistors

T_n and T_p	t_1	t_2
	millimicrosec	millimicrosec
2N501	25	34
2N559	27	27
SB240	33	27
OC170	30	190
OC44	210	210

The time t_1 is that taken to switch current out of the emitter of transistor T_p into that of T_n. Initially T_p is on, and the common emitter is established at 1·5 volts. When x falls in potential from +2 volts it can thus fall about 0·7 volt before the base-emitter diode of T_n starts to conduct. Similarly $\bar{x}$ has to rise from −0·5 to +1·5 volts before the transistor T_p attempts to switch off. Since the switching edges of x and $\bar{x}$ have a finite slope this latter action will take longer. Ultimately the effect of the change in levels of x and $\bar{x}$ will switch off T_p and turn transistor T_n on, thus establishing the common emitter at 0 volts. In order to achieve a small value for t_1 the hole storage of T_p must be small.

Now consider the reverse process when transistor T_n is to be switched off and T_p switched on. The common-emitter connection is at 0 volts. Thus x has only to rise 0·5 volt from −0·5 volt before T_n attempts to switch off. Similarly $\bar{x}$ ha only to drop 0·5 volt from +2 volts for the base-collector diod of T_p to conduct, thus attempting to switch the transistor on i the reverse mode.

In very asymmetrical transistors, those with a built-in dri field in the base region (e.g. OC170), the collector is a ver inefficient emitter. Thus no significant switching on of T occurs until the emitter has risen to 1·5 volts because the tran sistor T_n has switched off. In this instance t_2 will be greate than t_1.

The first three transistors listed in Table 4 have similar an fast switching times for both the t_1 and t_2 measurements. Th SB240 transistor is preferred because it is cheaper and it als possesses better d.c. characteristics which are discussed i Section 12.1.

The common-emitter connection of transistors T_n and T means that any attempt to modify the switching of one transisto also affects the other. Table 5 shows the effect on t_1 and t_2 c first reversing the emitter and collector of the transistor in th T_p position and also of a combination of different transisto types in these two positions.

Table 5

Switch Times of Various Combinations of the SB240 and 2N501 Transistors

T_n	T_p	t_1	t_2
		millimicrosec	millimicrosec
2N501	2N501	25	34
2N501	2N501*	32	24
SB240	SB240	33	27
SB240	SB240*	40	25
SB240	2N501	32	30
SB240	2N501*	36	28
2N501	SB240	28	24
2N501	SB240*	28	24

* Emitter and collector connections reversed.

The time t_2 is reduced by reversing the connections of T but the time t_1 increases and the maximum time of switching not improved. It is, however, reduced by certain combinatio of SB240 and 2N501 transistors.

Any two of the three transistors T, T′ and T″ in a stage of th carry path can switch together. The longest switch time of ar pair is the maximum switch time of the carry path. The mo satisfactory combination consists of three SB240 transistors co nected with common emitters as shown in Fig. 4.

(12.3) **Sum Logic**

The nature of the carry path demands that the carry pul must always be less than the amplitude of the function wav forms controlling the switch transistors. The function wav forms are conveniently at the levels which are standard elsewhe in the machine. Thus the carry signal has to be standardiz before the sum logic is performed. Alternatively it can be us in non-standard logical circuits to form the sum, which is the provided at standard levels for use elsewhere in the machine.

(12.3.1) **Sum Circuit of Experimental Adder.**

The carry path of the experimental adder described i Section 5 is shown in Fig. 4. The carry signal is taken fro the carry path by an emitter-follower circuit which gives a output pulse from +0·5 volt and −0·5 volt. This signal driv the base of a long-tailed-pair circuit as shown in Fig. 14.

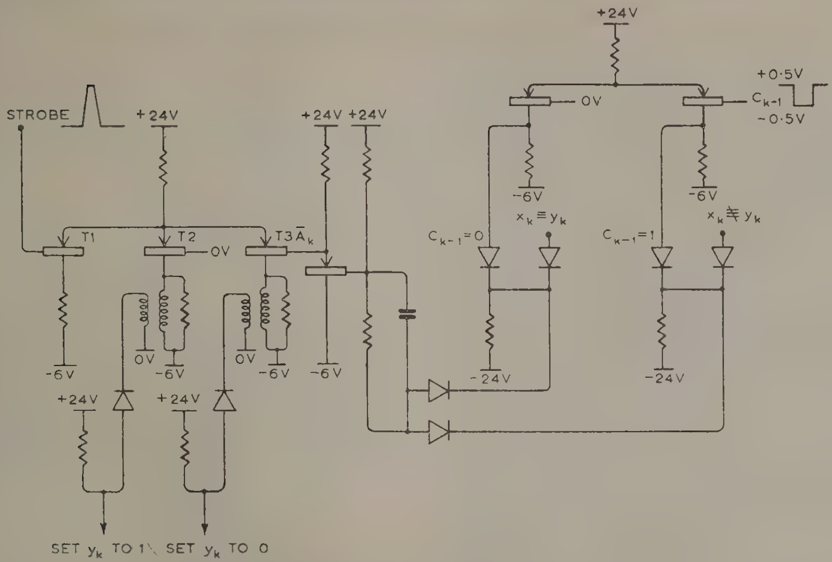

Fig. 14.—Sum circuit of experimental adder.

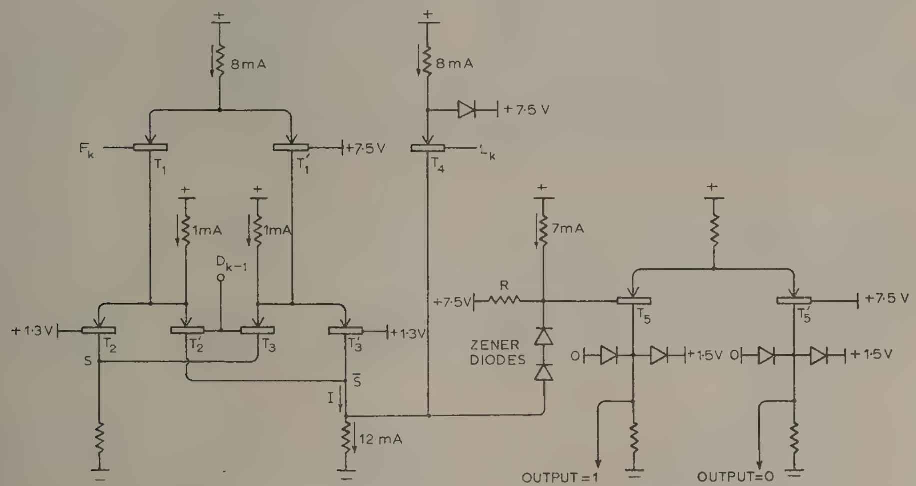

Fig. 15.—Sum/difference logic and output circuit of arithmetic unit.

Owing to the arrangement of the switch transistors on the arry path $+0{\cdot}5$ volt corresponds to carry $= 1$.

The outputs of the long-tailed pair are at standard level and eed diode gates which accept inputs $x_k \equiv y_k$ and $x_k \not\equiv y_k$ also t standard voltage levels.

The sum logic feeds the output circuit consisting of three ransistors T_1, T_2 and T_3. The signal $\bar{A}_k$ is at $-0{\cdot}5$ volt when he sum is zero. In the quiescent state the input to transistor T_1 s more negative than $-0{\cdot}5$ volt, so that this transistor is conlucting. When the strobe pulse occurs the base is taken more ositive than $+0{\cdot}5$ volt, so that one of the transistors T_2 or T_3 vill conduct depending upon the state of $\bar{A}_k$. This causes a ositive-going swing on the secondary of one of the transformers vhose primary is critically damped. The back swing, as the trobe pulse ends, sets the y_k flip-flop to the new sum. The trobe pulse is 100 millimicrosec wide and occurs at a minimum eriod of 200 millimicrosec, causing the adder to operate every ·2 microsec.

(12.3.2) Sum/Difference Circuit of Arithmetic Unit.

The sum/difference logic described in Section 6.2 is performed by the circuit of Fig. 15. The collectors of the transistors T_1 and T_1' form part of the 'long tails' of transistor pairs T_2, T_2' and T_3, T_3', respectively. The collectors of T_2', T_3' and T_4 are connected together and feed, through Zener diodes, the base of a standard long-tailed pair.

During arithmetic operations the base of T_4 is held above $+7{\cdot}5$ volts. The function waveform F_k controls the flow of current in T_1 and T_1', which after further control by the carry signal D_{k-1} can flow into the output circuit.

The current I is 9 mA when $F_k \equiv D_{k-1}$, which, in the case of addition, corresponds to

$$(x_k \equiv y_k) \text{ and } c_{k-1} \text{ OR } (x_k \not\equiv y_k) \text{ and } \bar{c}_{k-1}$$
$$\text{i.e. sum}_k = 1$$

alternatively $I = 1$ mA for $F_k \not\equiv D_{k-1}$, i.e. $\text{sum}_k = 0$.

The current in the common-collector resistor of T_2', T_3' and T_4 is defined at 12 mA. Thus when $I = 1$ mA the remaining 11 mA comes through the Zener diodes and 4 mA through the resistor R making the base of T_5 negative, with respect to $+7\cdot5$ volts, switching T_5 on and T_5' off. Conversely when $I = 9$ mA the surplus 4 mA flows into resistor R switching T_5 off and T_5' on.

During logical operation $I = 1$ mA and the transistor T_4 is switched by the function waveform L_k.

In the main accumulator gating of the logical function L_k is not required and the transistors T_4, T_5 and T_5' can be removed. Then both phases of sum output are available at S and $\bar{S}$, and can be connected to standard level by Zener diode circuits. The removal of T_5 and T_5' reduces the total delay through the adder by approximately 25 millimicrosec.

(12.4) List of Instructions

The following list of instructions can be obeyed by the arithmetic unit described in Section 6.

The B registers are either in a high-speed core store of 124 words or one of four fast flip-flop registers.

The S registers will receive their inputs from a fixed store of $0\cdot3$ microsec cycle time containing 4 000 words and a core store of $1\cdot7$ microsec cycle time holding 8 000 words.

List of Instructions Performed by Basic Arithmetic Unit.

Notation.

b, s = Contents of registers B and S before an instruction is obeyed.

b', s' = Contents of registers B and S after an instruction is obeyed.

0 $b' = s$
1 $s' = b$
2 $b' = b + s$
3 $s' = b + s$
4 $b' = b - s$
5 $s' = b - s$
6 $b' = s - b$
7 $s' = s - b$
8 $b' = b$ and s
9 $s' = b$ and s
10 $b' = b \not\equiv s$
11 $s' = b \not\equiv s$
12 $b' = b$ or s
13 $s' = b$ or s
14 $b' = b$ (with circular shift of one bit down) $+ s$
15 $b' = b$ (with circular shift of six bits up) $+ s$
16 $b' = b + s + 1$
17 $b' = - s$
18 New number to control if $b > 0$
19 New number to control if $b < 0$
20 New number to control if $b = 0$
21 New number to control if $b \neq 0$
22 New number to control if b is even
23 New number to control if b is odd

[The discussion on the above paper will be found on page 605.]

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FERRITE-CORE MEMORY SYSTEMS WITH RAPID CYCLE TIMES

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SUMMARY

Improvements in storage systems using currently available square-oop ferrite cores are considered. These enable the normal cycle me of 6–10 microsec to be reduced to less than 2 microsec. Effort as been concentrated on the word-selected two-core-per-digit arrange-ent, and the most promising techniques are those which involve artial-flux switching. A system is developed suitable for a store of 024 words of 52 digits with a cycle time of about 1·6 microsec. In a naller store of, say, 100 words, a cycle time of approximately ·6 microsec is feasible.

(1) INTRODUCTION

Square-loop ferrite cores are used almost exclusively in odern digital computers to provide the necessary large-capacity andom-access storage. They have proved extremely reliable in peration, and cycle times in the range 6–10 microsec are typical. t these speeds the limit to the maximum number of digits in single block of storage is largely economic, and whereas in a ritish computer there are typically 4×10^4 binary digits, in ne United States there is at least one store of $2{\cdot}5 \times 10^6$ digits nd many of 10^6 digits.

Some alternatives to the ferrite core have been developed ith the aim of simplifying the wiring and assembly of the orage elements. These are the Twistor[1] system and the multi-perture Rajchman plate system;[2] both of these have cycle mes of the same order as the simple ferrite-core systems. lthough the latter system is easy to wire, production of the late itself is difficult, since a large number of identical elements ave to be produced in a single process, and if any one element is nsatisfactory the whole plate is ruined.

Developments in logical design[3] and the advent of high-speed ansistors permit an arithmetic operation to be carried out in bout 1 microsec. If a machine is to take full advantage of such peeds, its storage system must have a cycle time of approxi-nately ½–1 microsec. Proposals for such systems which are in ne early development stage include the use of thin magnetic[4, 5] lms and superconducting devices.[6, 7] Neither of these systems in a sufficiently advanced state to be included in a machine hich has to be in production in two to three years' time. How-ver, preliminary experiments[8] on cycle time are very promising nd indicate that a figure of 100 millimicrosec might well be chieved.

A computer operating normally takes its orders from sequential ddresses in the main store. An effective reduction in the access me to these orders can be achieved by providing a number of iscrete storage blocks and arranging that sequential addresses pply to each of these blocks in turn. For example, if a computer as four blocks of storage each with a cycle time of 6 microsec, nen, depending on the sequence of demanded addresses, the fective access time to information in the complete store could ave any value between 1·5 and 6 microsec. Departure from ne smaller figure occurs because the numbers demanded as a result of each order may not be addressed in any particular sequence and large changes of address also occur during control transfers. Use of this technique necessitates a further reduction of only 2 or 3 times in the cycle time in order to satisfy the requirements previously stated. Some or all of this factor may yet be realized from developments in ferrite materials (Section 9.1.1), but to achieve an immediate solution in terms of the ferrite cores now available, a different approach has to be adopted. The paper describes a system suitable for providing 1 024 words each of 52 digits in a single storage stack with a cycle time better than 2 microsec.

The provision of B-registers in a high-speed parallel computer is also a problem. These could be realized in terms of fast transistor flip-flops, but this technique is expensive and sets an economic limit on the number of such registers. From the user viewpoint the provision, for example, of 32–128 B-registers would be an advantage in order to assist autocode procedures and time-sharing techniques among a number of peripheral equipments. It is thus desirable to have store of 128 words maximum with a cycle time not greater than 0·6 microsec. A system capable of this performance is also described.

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Prof. Kilburn is Professor of Computer Engineering, Manchester University.

(2) USE OF SQUARE-LOOP CORES AS STORAGE ELEMENTS

The use of square-loop ferrite cores as binary storage elements is based on the fact that the material of the cores exhibits at least two distinct states of remanent flux. A typical B/H loop of a 1·3 mm core, obtained by cycling the core between the limits $\pm H_f$ is shown in Fig. 1. This major loop does not

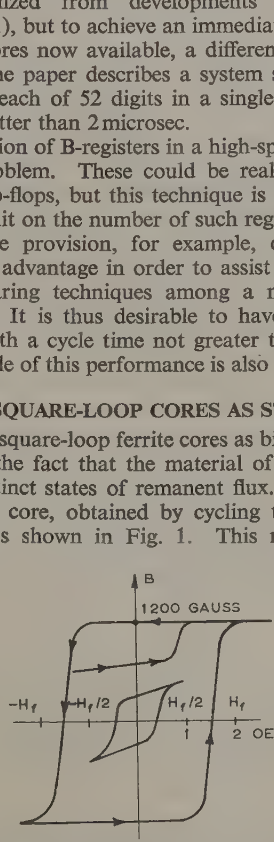


Fig. 1.—Major and minor hysteresis loops.

represent the full saturation limits of the material, and other loops do exist concentric with it. In some applications flux changes significantly less than normal occur, and two types of minor loops are indicated in the Figure.

When a core is driven along the relatively flat portions of the hysteresis loop the rate of change of flux is proportional to the rate of change of current, i.e. the voltage developed across a 'read' winding linking the core is given by $L\,di/dt$. The magni-tude of L (for a given direction of current) depends on the remanent state of the core, since the slopes of the upper and lower saturation limits in the region of the zero field axis are rather different. This fact has been used as the basis of a method of

non-destructive reading of cores; but there are difficulties due to the small signal level, and a rather low discrimination ratio (about 2 : 1) is obtained.

If a current step is applied to a core so as to switch it from one remanent state to the other it is apparent from the output voltage waveform that flux continues to change in the core after the drive current has attained a steady value. Under these circumstances the core does not behave in an inductive fashion but more as a resistance, the value of which is time dependent. It has been shown, both theoretically and experimentally,[9, 10] that the duration of the output pulse, T_s, may be related to the applied m.m.f. by the expression

$$(I - I_c)T_s = K \text{ or alternatively } (H - H_c)T_s = K' \quad . \quad (1)$$

where H_c is the coercive force which in a typical coincident current storage system lies midway between $\frac{1}{2} H_f$ and H_f

i.e. $$H_c = \tfrac{3}{4}H_f \text{ or } I_c = \tfrac{3}{4}I_f \quad . \quad . \quad . \quad . \quad (2)$$

Eqn. (1) is obeyed closely for values of I between about $1 \cdot 5\, I_c$ and $3\, I_c$. Outside this range faster switching than predicted by eqn. (1) is obtained, because for $I < 1 \cdot 5\, I_c$ full flux switching does not occur, and at larger values the mode of switching changes.[11] The form of eqn. (1) indicates that the switching time of a core may be reduced by increasing the drive current; this is a technique which is particularly applicable to word-selected parallel stores where the reading current is not limited by selection requirements as in the half-current selection system. The results of switching measurements on a number of square-loop cores are given in Section 9.1.1.

Switching these cores at a high repetition frequency causes a marked increase in the temperature of the core since the thermal conductivity of the material is low. Measurements on cores at different temperatures are given in Section 9.1.5, and these show that as the temperature of the core increases:

(a) H_c is reduced, and thus the core tends to switch faster for a given drive.
(b) The flux density decreases.
(c) The slope of the relatively flat portions of the loop becomes more pronounced and therefore the voltage output from a core 'disturbed' by a half-current pulse is temperature dependent. At the Curie temperature, which varies with different ferrites between 100 and 300° C, the core loses all its magnetic properties, although these return as it cools down. In general, ferrites with the higher coercivity have a higher Curie point.

There are two basic ways in which cores can be arranged to form a storage system, known as the coincident-current and word-selection systems, respectively. Both these systems can be modified to improve their speed of operation, but the first system lends itself to serial[12] and the second to parallel operation. The discussion is therefore confined to the latter system since parallel operation is required.

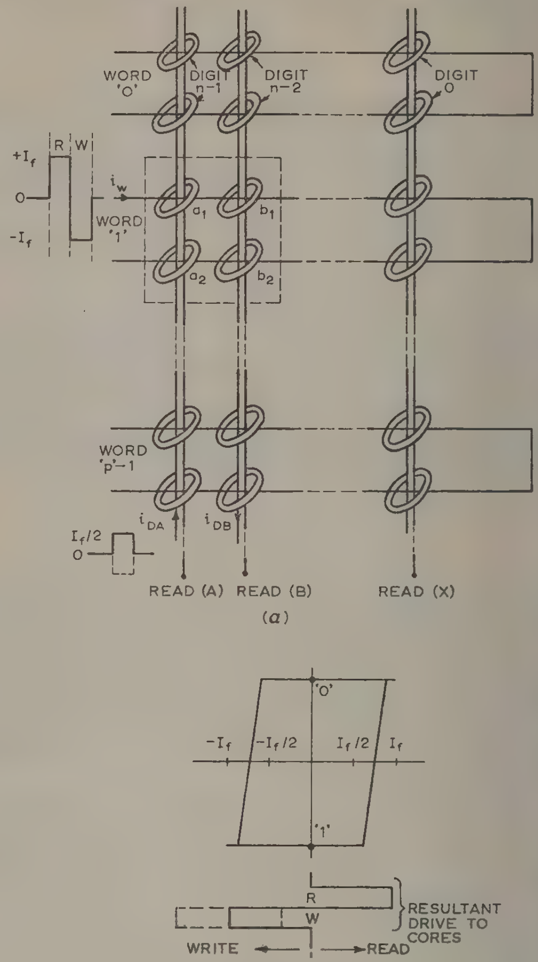
(3) BASIC SYSTEM USING TWO CORES FOR EACH DIGIT

The system is shown in Fig. 2, which illustrates the method of wiring and also the necessary current waveforms. The cores of a core pair representing a given digit are normally in opposite magnetic states, and to explain the system it is assumed that initially

Core a_1 is in the '1' state
Core a_2 is in the '0' state } digit A

Core b_1 is in the '0' state
Core b_2 is in the '1' state } digit B

The 'read' current is defined as positive on the idealized loop of Fig. 2(b). The state of the digit is defined by the state of the a_1, b_1, c_1, etc., cores, and thus digit $A = 1$ and digit $B = 0$.



(b)

Fig. 2.—Basic system.
(a) Wiring layout.
(b) 'Read'/'Write' cycle.

When the word is selected a current $i_w = + I_f$ flows in the word wire for a time termed the 'read' period. During this time all the cores in the selected word are returned to the '0' state, i.e. in digits A and B cores a_1 and b_2, respectively, change their state. There are also small reversible flux changes in the cores a_2 and b_1 which are driven along the nominally flat portions of the hysteresis loop. The 'read' wire for each digit position links each core of a core pair in an opposite sense with respect to the 'read' current, and thus the output signal represents the difference in the rates of change of flux in the cores of a core pair. The polarity of the 'read' pulse depends on which core of the core pair reverses its flux state in the 'read' period; if digit $A = 1$ is indicated by a negative pulse then that from digit $B = 0$ will be positive.

The reading process destroys the stored information, and it is necessary to rewrite if it is to be retained. A 'write' (or 'rewrite') period always follows a 'read' period. In the 'write' period the word current, i_w, is reversed and assumes a value $-I_f$. Simultaneously currents i_{DA}, i_{DB}, etc., of value $\pm\frac{1}{2}I_f$ are applied to the digit wires. These assist the writing action of the 'word' current in one core of a core pair and inhibit its action in the other. The sign of the digit current is determined by the infor-

nation to be written, and if digits A and B are to be rewritten n their original form the directions of the currents i_{DA} and $_{DB}$ will be negative and positive, respectively. The duration of he 'read' period is defined by the switching time of the core ppropriate to a drive of $+I_f$ and of the 'write' period to a lrive of $-\frac{3}{2}I_f$. These switching times set the minimum cycle ime of the system. In practice the cycle time will also be nfluenced by the type of selection system and driving circuits ised.

The use of two cores per digit has disadvantages in that the hysical size and cost of the store are increased relative to a ne-core-per-digit system. However, there are a number of ignificant advantages which make the system worth while. These are as follows:

(*a*) The back-e.m.f. developed across the selected 'word' wire luring the 'read' and 'write' phases of the store cycle is independent of the digit pattern of the word, thus easing the design roblems associated with the word-selection matrix.

(*b*) The binary states of the stored digits can, in some arrangenents, be indicated by signals of opposite polarity, giving mproved discrimination over the 'signal/no-signal' conditions btainable from one-core-per-digit systems.

(*c*) Owing to imperfections in the word-selection matrix, ignals can arise from partially selected words which interfere vith the 'read' output from the selected word. In two-coreer-digit systems these interference signals are smaller than n a one-core-per-digit system.

(*d*) There is the possibility of developing partial flux-switching ystems. In such systems one state of saturation of the cores is sed as a reference and both cores are set into this state in the read' phase of the store cycle. During the 'write' phase flux hanges occur in both cores, but it is arranged that one core hanges more than its partner and also that the larger flux hange is only about 25% of a complete switch from one saturaion limit to the other. The 'read' output is due to the difference n flux content of both cores and thus is not affected by the bsolute value of flux in either core. This technique permits igh operating speeds to be associated with reasonable power equirements.

(3.1) A Coded System

Operation of the coded system is explained by taking an xample in which the cores of a word are divided into groups of our. In Fig. 2(*a*) such a group is formed by the cores a_1, a_2, $_1$, b_2. Only one of these cores is normally set to '1', the others emaining in the '0' state. The position of the core in the '1' tate determines the value of the two digits, i.e. A and B. The etailed coding is shown in Table 1.

A comparison of the system with the conventional one shows hat the number of cores changing state at any one time is educed by a factor of 2. This has the advantage of first making provision of the word current simpler, and secondly, the number of digit-current drivers required to operate in the 'write' period is halved.

Table 1

CODED SYSTEM

Digit A	Digit B	Cores				'Read' wire output	
		a_1	a_2	b_1	b_2	A	B
0	0	1	0	0	0	—	Zero
0	1	0	0	0	1	Zero	+
1	0	0	0	1	0	Zero	—
1	1	0	1	0	0	+	Zero

The discrimination ratio on the 'read' signal is reduced since on any 'read' wire the output signal can be positive, negative or zero. In the 'rewrite' period these three levels of output cause digit currents $+\frac{1}{2}I_f$, $-\frac{1}{2}I_f$ or zero to be applied to the digit wires. In order that the computer and the store can communicate with each other, some additional gating is necessary to convert from the conventional system of digit representation to or from the coded form.

The principle of this coded system may be extended so that, in general, 2^n cores represent n digits. For $n > 2$ the number of cores required exceeds two per digit, and thus the capital cost is relatively greater, although the power saving increases in proportion to n. If the increased cost of the system for $n > 2$ is acceptable, the limit to n is set by the increasing complexity of the coding conversion equipment.

(3.2) Word-Selection Matrix

The word-selection matrix may consist of diodes, either alone or in combination with linear current transformers, or transformers wound on square-loop cores. Diodes capable of transmitting current pulses of the order of 500 mA at the repetition frequencies proposed are not readily available. In a system, two diode matrices would be necessary to provide the bidirectional word current, and the address staticizor and decoding mechanism of the store would have to remain set during the whole store cycle, thus making it impossible to overlap the set time of the address staticizors and address decoding time into the 'write' period. This may be overcome at the expense of providing a separate address staticizor and decoder for the 'write' current matrix. It is possible, when diode matrices are used, to have an unbalanced word current, i.e. the product of the 'read' current amplitude and its duration need not equal the product of the 'write' current amplitude and duration, as is inevitably the case if an a.c. coupling is used to provide the word current. Diodes may also be combined with linear transformers, and here the diode current can be limited to the primary current of the transformer. This is not an attractive alternative when transistor drive circuits are used, as the permissible collector voltages are usually low and the available currents high.

A third alternative is to use transformers wound on square-loop cores arranged in a matrix and selected by coincident-current techniques. In a typical arrangement the switch cores are saturated in a given direction by a d.c. bias which is common to all switch cores. Then one switch core is selected by a coincidence of currents on the X and Y wires of the matrix, which overcome the bias and provide the 'read' current. When the drive currents are removed the switch core relapses to its original state under the influence of the d.c. bias. This provides the 'write' current automatically, making it unnecessary to keep the address staticizor and decoder set during the 'write' period. However, switch cores are very inefficient current transformers since a large proportion of the primary m.m.f. is absorbed as magnetizing current; this is the price paid for the inherent memory of the switch cores.

(3.3) Modifications to Increase the Speed of the Basic System

The minimum cycle time of the basic system is limited by the switching time of the cores in the 'read' and 'write' periods. The switching time in the 'read' period may be reduced by increasing the 'read' current, but the magnitude of the 'write' current is limited since its action must be inhibited where required by the digit current, and the magnitude of this current

must not exceed $\frac{1}{2}I_f$ since it passes through the cores of all the unselected words.

(3.3.1) **Illinois System** (see Reference 13).

A modification to the basic system which is faster and yet retains a balanced word-current waveform is the Illinois system. The arrangement is similar to that shown in Fig. 2, but the word current is $+\frac{3}{2}I_f$ in the 'read' period followed by $-I_f$ in the 'write' period. The minimum duration of the 'read' pulse, t_1, is sufficient to allow the cores to switch with the applied drive, and the duration of the 'write' current is then $\frac{3}{2}t_1$. The word-current waveform is thus balanced and the minimum cycle time is $\frac{5}{2}t_1$. If an unbalanced word-current waveform could be accepted the 'write' period could also be of duration t_1 and the cycle time would become $2t_1$.

(3.3.2) **Bias System A.**

The limitation imposed on the digit currents inherent in the two systems previously described may be relaxed by suitable biasing of the memory cores. Several systems are possible using this technique; the first to be described is called 'bias system A'.

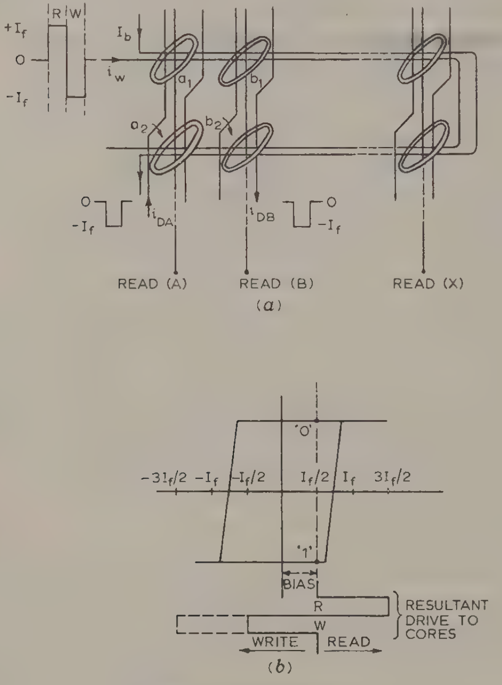


Fig. 3.—Bias systems.
(a) Wiring layout.
(b) Operation of system A.

The wiring diagram is now that of Fig. 3(a). The cores on word wire are given a d.c. bias $I_b = +\frac{1}{2}I_f$, and the 'read' current is $+I_f$. These provide a net reading current of $+\frac{3}{2}I_f$. In the 'write' period the word current becomes $-I_f$, and the action of this current is assisted [see Fig. 3(b)] by a digit current $-I_f$, which flows simultaneously in the appropriate digit wire. The net 'writing' current is $-\frac{3}{2}I_f$ and the minimum cycle time is $2t_1$ (where t_1 is the switching time of the cores for a drive of $\frac{3}{2}I_f$.) A balanced word-current waveform is retained. The digit currents always act in opposition to the bias current and thus will cause only a relatively small disturbance in the cores of unselected words.

At first sight this system has two unattractive features:

(a) The digit wires thread one core of a core pair and the wiring of the matrix as shown in Fig. 3(a) is relatively complex and would be expensive to produce.
(b) The digit currents are twice as large as in previous systems.

Both of these disadvantages may be overcome simultaneously (in a large store at least) by arranging that the cores a_1, b_1, c_1, etc., are in one core plane and the other cores of the word, a_2, b_2, c_2, etc., are in another plane mounted back to back with the first. By this means the two digit wires of a given core pair are in separate planes and these thread every core in a given digit position. The wiring of the planes is then a fairly simple procedure. In a large store the length of the digit wire is a limiting factor in the design of the digit drivers, this being particularly important when transistors are to be used since the maximum permissible collector voltage tends to be low. In systems where the digit current is of magnitude $\frac{1}{2}I_f$ the digit wire threads $2N$ cores, where N is the number of words in the store, and is therefore twice as long as the digit wire in the bias system which threads only N cores, although here the digit current is of magnitude I_f. Thus the voltage required to establish the digit currents at a given rate will be the same in either case, assuming that the digit wire is mainly inductive.

If the core planes are arranged in this way one half of the 'read' wire lies in each plane, each half threading N cores. If the 'read' wire is now earthed in the middle a push-pull output is obtained at the free ends and a difference amplifier may be used in the 'read' circuits. The transmission delay on the 'read' signal from the cores in a given word will then be only half as long as that obtained in the first two systems described.

(3.3.3) **Bias System B.**

The bias system described may be further modified to reduce the minimum cycle time by increasing the magnitude of the 'read' current. This modification is called 'bias system B'. In this system the 'read' current is increased to $\frac{3}{2}I_f$ and its duration can therefore be reduced to $\frac{2}{3}t_1$. In the 'write' period the word current is $-I_f$ and its duration is t_1. The word-current waveform is still balanced, and the minimum cycle time has been reduced to $1\frac{2}{3}t_1$.

Further modifications to bias system A are possible which involve partial flux switching and result in either a further reduction in the minimum cycle time and/or a relaxation on driving requirements of the system. These are considered in a later Section. The systems discussed in this Section are compared under various headings in Table 2.

(3.3.4) **A Comparison of Systems.**

In Table 2 the 'read' and 'write' switching times are calculated from eqn. (1) and are for guidance only. It is assumed that eqn. (2) is true for the cores used. In the column of minimum cycle times expression (i) applies when unbalance in the word-current waveform can be tolerated; these are modified to (ii) when the word-current waveform must be balanced. For bias system A the 'read' and 'write' currents are of equal magnitude and so are the 'read' and 'write' switching times. Thus there is no object in making the word-current waveform unbalanced.

The quoted times are for the F764/S4 memory core, which has the following characteristics:

$$I_f = 500\,\text{mA in one turn}$$
$$I_c = 380\,\text{mA in one turn}$$
$$K = 160\,\text{mA in one turn-microsec}$$
$$\frac{K}{I_c} = 0\cdot 42\,\text{microsec}$$

Table 2

COMPARISON OF SYSTEMS

System	Net word current	Digit current	'Read' switching time t_r	'Write' switching time t_w	Minimum cycle time	Notes
Basic system	$+I_f$, 0, $-I_f$	$\pm\frac{I_f}{2}$	$\frac{K}{I_f - I_c} = \frac{3K}{I_c}$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	(i) $4\cdot0\frac{K}{I_c} = 1\cdot68$ microsec (ii) $6\frac{K}{I_c} = 2\cdot56$ microsec	Digit current inhibits action of 'write' current where switching is not required and assists it where switching is required.
Illinois system	$+\frac{3}{2}I_f$, 0, $-I_f$	$\pm\frac{I_f}{2}$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	(i) $2\cdot0\frac{K}{I_c} = 0\cdot84$ microsec (ii) $2\cdot5\frac{K}{I_c} = 1\cdot05$ microsec	Similar to basic system, except 'read' current increased.
Bias system A	$+\frac{3}{2}I_f$, -0, $-I_f/2$	$-I_f$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	(i) $2\cdot0\frac{K}{I_c} = 0\cdot84$ microsec	Memory cores biased to $+\frac{1}{2}I_f$, digit current acts in opposition to bias current and assists on action of 'write' current where required.
Bias system B	$+2I_f$, 0, $-I_f/2$	$-I_f$	$\frac{K}{I_f - I_c} = 0\cdot6\frac{K}{I_c}$	$\frac{K}{\frac{3I_f}{2} - I_c} = \frac{K}{I_c}$	(i) $1\cdot6\frac{K}{I_c} = 0\cdot67$ microsec (ii) $1\cdot67\frac{K}{I_c} = 0\cdot71$ microsec	Similar to bias system A; except for increased 'read' current. In this case the minimum cycle time with a balanced word current waveform is limited by the 'write' switching time since here $t_w > 1\cdot5t_r$.

These are calculated using eqn. (1). More accurate values can be obtained from the curves of Fig. 13. The minimum cycle times quoted are theoretical, and in practice these figures will be increased owing to the time needed for establishing the operating currents. At present, this latter figure is estimated at 0·6 microsec, and thus bias system A, for example, should cycle in the order of 1·5 microsec.

(4) EXPERIMENTS ON BIAS SYSTEM A

The experiments on bias system A have been carried out on a single word driven from a switch core. The design of a suitable switch core and a method of estimating the required primary m.m.f. are considered in Section 9.2. The experimental arrangement is illustrated in Fig. 4, which shows the word wire threading 104 cores corresponding to 52 digits. The resistance R_s has been included to reduce the time-constant associated with the secondary circuit of the switch core. The memory cores are biased by a current of +250 mA which flows in a 'bias' wire not actually shown in the Figure.

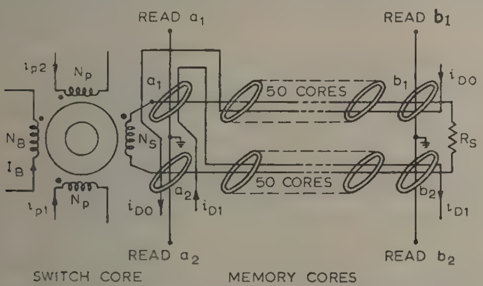


Fig. 4.—Bias system A. Experimental arrangement.
The bias wire on the memory cores is not shown.

(4.1) Flux-Change Experiment

The switch core type F00 2/S4(L) is biased to $-2\cdot43$ AT (as indicated by calculations in Section 9.2.3) and pulsed against this bias with currents i_{p1} and i_{p2} in the primary windings of N_p turns. Half of the 104 memory cores on the 'word' wire are supplied with a digit pulse i_{D0} of 500 mA, and the flux change in these cores during the 'read' period is measured as a function of the switch-core primary m.m.f. for various values of R_s. The results are plotted in Fig. 5, which also shows the pulse configuration used. The vertical scale is proportional to flux changed in the memory cores, this being plotted as a percentage of the normal flux change obtained when the cores are switched from +250 to +500 mAT.

When the saturation limits of the system are reached, measurements indicate that the switch core, which is traversing a loop very near that of maximum saturation, is changing approximately 1·2 times as much flux as would be obtained by the normal switching of 52 memory cores, i.e. some of the available flux change is lost in the transfer. The voltage equivalent to this lost flux is developed across the resistance and stray inductance of the word wire loop. By comparing the sum of the voltages developed across the switched cores, the unswitched cores and R_s with that developed at the switch-core secondary it is possible to estimate the effect of the stray inductance ($L_s \simeq 1\,\mu$H) on the efficiency of flux transfer. This experiment indicates that R_s is the dominating factor.

Fig. 6(*a*) shows the waveforms obtained under the conditions of the experiment with $R_s = 2$ ohms and a total primary drive of 4·8 AT. The 'read' time is about 0·4 microsec and the 'write' time 0·45 microsec. It is evident that more flux change occurs in the 'write' period than is obtained in the 'read' period, the difference in the two being made up when the 'write' and digit currents are removed. The flux waveform indicates that the core operates on a loop such as that of Fig. 6(*b*). In the

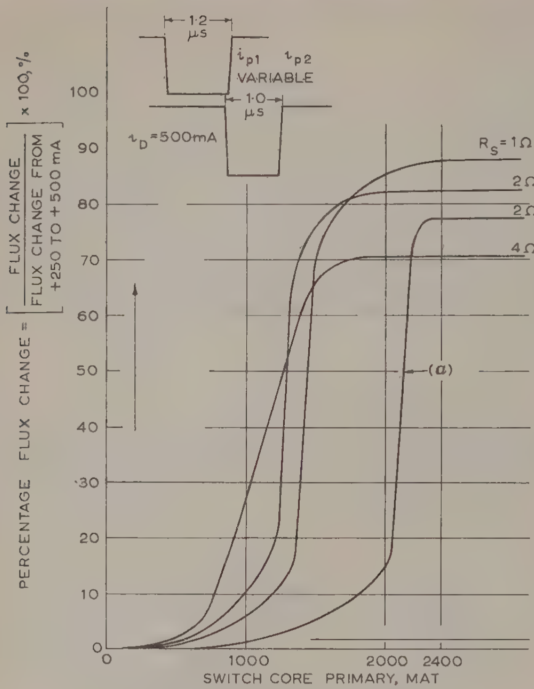


Fig. 5.—Flux-change experiment.

Pulse rise and fall times are about 50 millimicrosec.
$N_p = 12$ turns $N_B = 3$ turns $N_s = 2$ turns
$I_B = 810$mA $I_b = 250$mA.

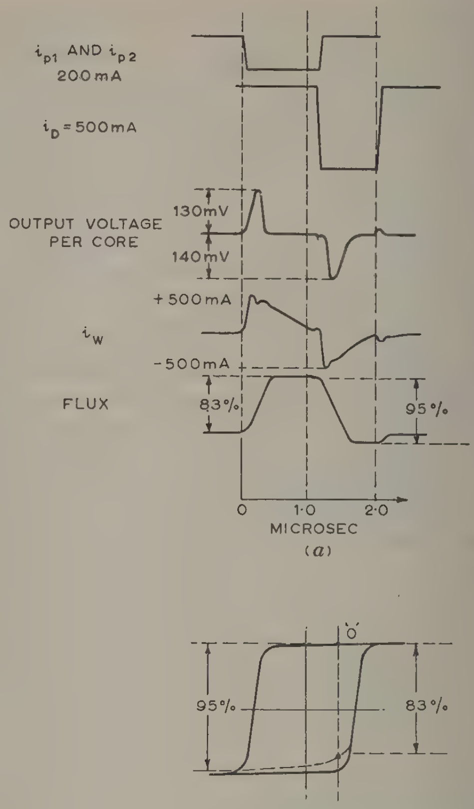


Fig. 6.—Flux-change experiment.

(a) Waveforms.
(b) Operating loop.

'read' period the available flux change is limited by the upper saturation limit of the core. This has been checked by pulsing the cores between successive 'read'/'write' cycles with a current pulse of variable amplitude in the direction of the bias. Those cores in the '0' state at the end of the 'write' cycle show a small reversible flux change, indicating that these cores are saturated, while the result of pulsing those cores left in the '1' state shows that the knee of the operating loop is situated at about +300 mAT.

The curves of Fig. 5 show that the approximate analysis of Section 9.2 is justified, but the presence of inductance in the word wire loop makes it necessary to include R_s in order that the 'write' current can decay sufficiently quickly when the writing action is complete. The effect of R_s on the efficiency of flux transfer is largely compensated by the excess flux content of the switch core, and the system achieves an approximate full-flux switch ($R_s = 2$ ohms). The current pulses used in the experiment are rather wider than necessary, and Fig. 6(*a*) indicates that an active duration of 0·5 microsec should be adequate.

(4.1.1) Operation under Computer Conditions.

In Fig. 7 the waveforms obtained when the active duration of the current pulses is reduced to 0·5 microsec are shown; the rise and fall times are arranged to be 0.2 microsec to simulate the limitations of a practical system. The rise and fall times of the digit and primary currents may be overlapped and a cycle time of about 1·6 microsec is indicated. It is found that the current-discrimination properties of the system become more pronounced as the active duration of the primary currents is reduced. The duration of the current pulses is then too short to allow a significant flux change to occur in the memory cores until the current amplitude approaches that required to reverse the flux state of the switch core completely. This is illustrated in Fig. 5, the curve marked (*a*) being taken with the pulse configuration of Fig. 7. The curve also shows that 5% less flux is changed under these conditions than previously, this being due to pulse-width limitation, and it indicates that the duration of the current pulses should not be less than the 0·5 microsec used.

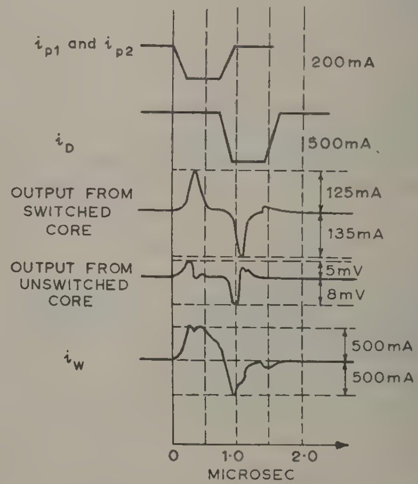


Fig. 7.—Waveforms obtained under computer conditions.

The system approximates very closely to a 'full flux switching' system, a flux change of 78% being obtained in the 'read' period, and further tests have been carried out to assess its behaviour under conditions likely to arise in a computer at a cycle time of 1·6 microsec.

The sensitivity of the operating loop of the memory cores, which may be in either the '1' or '0' state, to repeated digit pulses has been investigated. Digit pulses of 550 mA amplitude (10% greater than the nominal 500 mA) were allowed to 'disturb' the cores between successive 'read'/'rewrite' cycles. Those cores left in the '0' state after the writing operations were most affected. This is to be expected since the action of the digit pulse in this case is to move the core away from saturation. The cores left in the '1' state after the writing operation are relatively unaffected, since the tendency is to increase the state of saturation of the core. For cores in the '0' state the first 'disturb' pulse results in a small (approximately 10%) increase in the amplitude of the next 'read' signal. This increase is about 12% after the core has been disturbed three or four times and subsequently remains unchanged for any further pulses. There is no appreciable change in the 'read' output from a disturbed core in the '1' state for any number of 'disturb' pulses.

In another experiment the digit currents i_{D0} and i_{D1} [Fig. 4] were so arranged that a given core was written into the opposite state to that indicated in the 'read' period. This test indicates whether a core which is switched from the '1' state to the '0' state during the 'read' period is, in fact, in an identical magnetic state to that of a core which is switched from the '0' state and left in that state at the end of the 'read' period. Within the limits of experimental accuracy this is found to be the case here.

If the store is cycled at a high repetition frequency equivalent to a cycle time of 1·6 microsec, both the switched memory cores and the switch core suffer an appreciable rise in temperature. In this experiment the switch core was immersed in an oil bath to improve the efficiency of heat dissipation but the memory cores were in free air. It was found that the system operated satisfactorily at this frequency, the rise in temperature of the memory cores to almost 45°C resulting in an approximately 15% increase in the amplitude of the 'read' output voltage, and a slightly reduced switching time. The experiment indicates that some temperature control of the system would be essential in practice. The thermal time-constant of the system is of the order of 2 sec.

It is concluded that the system described in this Section represents a possible solution to the problem of providing a large-capacity store with a cycle time of less than 2·0 microsec. Alternative schemes which involve partial flux switching are now considered.

(5) PARTIAL FLUX SWITCHING

(5.1) Basic System

The normal flux change occurring in memory cores can be restricted by limiting the available flux from the word-driving switch core. Hence the F002/S4 (L) switch core used previously (see Fig. 4) has been replaced by a shorter core of similar radius, namely F002/S4(S).

With these conditions the peak amplitude of the 'read' output signal from the cores of a core pair has been plotted against digit current i_D and the results and drive waveforms used are shown in Fig. 8. Throughout this experiment the relationship $i_D = 2I_b$ was maintained and the word-current amplitude kept constant at ± 500 mA. The maximum difference signal occurs at a digit current of -200 mA, and this signal (approximately 30 mV) constitutes the 'read' output voltage of the system. The waveforms obtained at this optimum value of digit current are shown in Fig. 9(*a*).

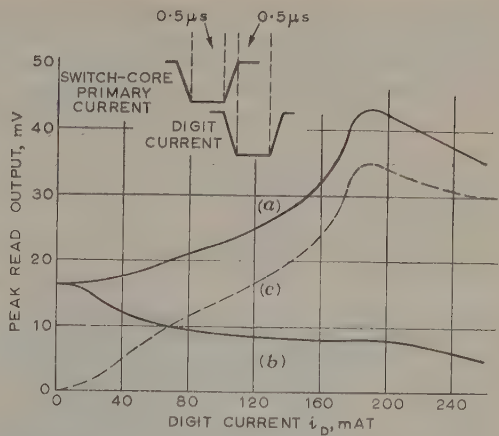


Fig. 8.—Variation of 'read' signal with digit and bias currents.

Word current = ± 500 mA.
Bias current, $I_b = +i_D/2$.
(*a*) Core with digit current.
(*b*) Core without digit current.
(*c*) Difference signal.

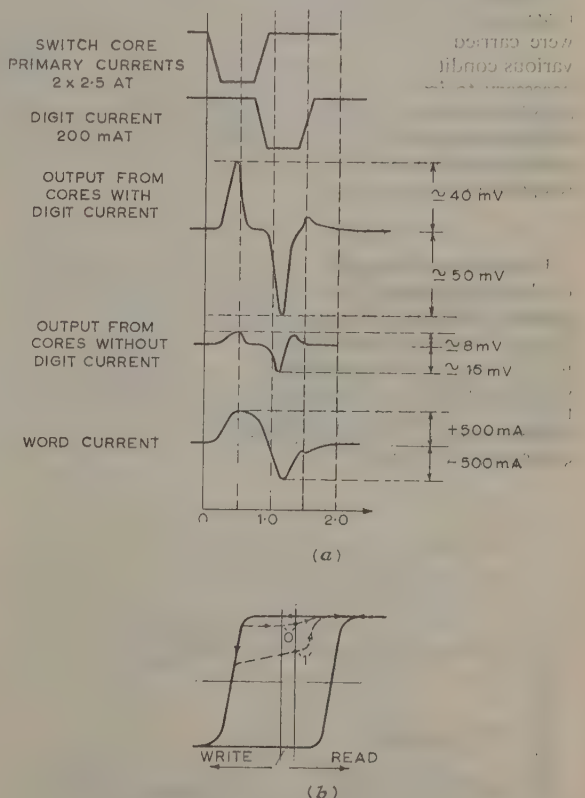


Fig. 9.—Partial flux switching.

(*a*) Waveforms.
(*b*) Operating loops.

The operative B/H loop of the memory cores is indicated in Fig. 9(*b*). The flux change in the 'read' period (26% of normal for a '1') is restricted by the saturation limit of the core material,

and in the 'write' period by the amplitude and duration of the net switching current (38% of normal for a '1'). Those cores in the '0' state before the 'read' period and remaining in the '0' state after the 'write' period are only subject to the action of the word and bias currents and the flux change is about 5% of normal.

Tests with a current pulse of variable amplitude applied in the same direction as the bias to the cores between successive 'read'/'write' cycles show that:

(a) The '0' state is situated very near the upper saturation limit of the operating loop.
(b) The '1' state is situated some 50 mAT away from the knee.

The system described is analogous to bias system A except that a smaller flux change is used. The speed of the two systems is the same, but the power requirements of the partial flux switching system are considerably smaller for three reasons:

(a) The digit current has been reduced from 500 to 200 mA.
(b) The voltage appearing across the word wire is reduced by a factor of about four, though the primary m.m.f. requirements of the two systems are much the same, because the radii of the switch cores and the required switching times are similar.
(c) Since a shorter switch core is used, the inductance of the half-selected switch cores will be less and the switch-core matrix will be easier to drive.

This system is therefore preferred to bias system A for the large-capacity store. Tests similar to those in Section 4.1.1 were carried out to check the operation of the system under various conditions. These proved satisfactory, but it was found necessary to immerse the memory cores as well as the switch core in an oil bath. This would indicate that the operating loop of the memory core is more sensitive to temperature changes under partial flux switching conditions despite the fact that the temperature rise is smaller.

(5.2) Improvement in Speed

Improvement in speed is obtained by increasing the amplitude of the word current. For experimental purposes where it was necessary to vary this current, it was convenient to supply it from a linear transformer and its approximate form is shown in Fig. 10. In the experiments described the duration of the operating pulses was maintained at 0·3 microsec since a cycle time of 0·6 microsec is required. The memory cores were immersed in an oil bath to achieve adequate cooling and thus satisfactory operation.

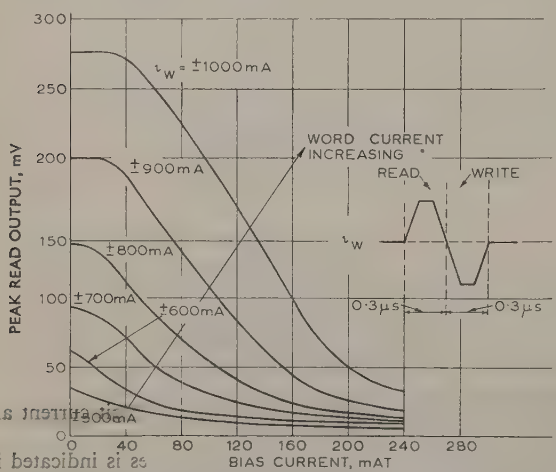


Fig. 10.—Variation of peak 'read' output voltage with bias current.

(5.2.1) Effect of the Bias Current.

There are three variables to be considered, the amplitudes of the word current i_W, the digit current i_D and the bias current I_b. In any cycle of the store, one core of each core pair is subject to the action of the word and bias currents only, and this core must not experience a significant flux change in the 'write' period. Fig. 10 summarizes the way in which the action of the word current on these cores is modified by the bias current. The 'read' output signal decreases as the bias is increased despite the net increase in 'read' current, since the flux change is limited by the saturation characteristic of the core and by flux change in the preceding 'write' period. Since the effective 'write' current decreases with bias this latter flux change will decrease in magnitude. It is observed that the flux change in the 'write' period is greater than that in the 'read' period, and the difference is made up after the word current is removed, when the core is acted upon by the bias current alone.

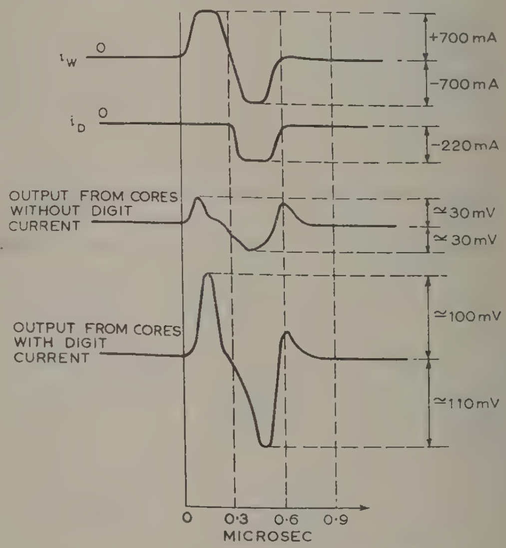


Fig. 11.—Waveforms. A faster system.

A typical voltage waveform is shown in Fig. 11, which shows that the recovery of the core back to its original state on the removal of the 'word' current is relatively fast. Thus the bias current has the effect of stabilizing the quiescent operating point on the B/H characteristic. The recovery of the cores back to a stable position when the 'write' current is removed does not limit the maximum operating speed of the system significantly. This is discussed in Section 5.2.4.

(5.2.2) Effect of the Digit Current.

The curves of Fig. 10 suggest two possible modes of operation:

(a) A system without bias in which the digit current is used to reduce the switching effect of the 'write' current where required.
(b) A system with bias in which the action of the bias is inhibited by the digit current where required.

The first system appears to be more attractive since Fig. 10 could be interpreted to state that a digit current of the order of 100 mA gives adequate discrimination between the output signals from the cores of a core pair. However, this is not found to be so in practice, and digit pulses of the order of 200 mA are found to be necessary, showing that a pulse is not so powerful as a d.c. bias of comparable magnitude in inhibiting the action of the

'write' current (see Section 9.1). Moreover, the stability of the operating loops of this mode of operation was found to be very sensitive to repeated digit pulses, the direction of which was such as to give reduction in subsequent 'read out' signals from the cores. This system has been rejected for these reasons:

In the second system the digit current is used to inhibit the action of the bias current and to assist the switching action of the word current in the 'write' period. It is found that if the magnitude of the digit current is twice that of the bias current, so that the net 'read' and 'write' currents are equal in magnitude, the 'read' output voltage from the core is almost equal to that obtained when the core is cycled with the word current alone. The validity of this statement is limited to bias currents less than the knee of the operating loop. With these observations in mind, values of i_w, I_b and i_D may be chosen by reference to Fig. 10.

A suitable choice for the word current is $i_w = \pm 700$ mAT, which, when used with a bias of +100 mAT and a digit current of the order of −200 mAT will give a difference of about 3 : 1 in the 'read' output signals from the cores of a core pair. The variation in the peak 'read' output voltage with digit current under the specified conditions and word and bias currents is shown in Fig. 12(*a*), where it is seen that the actual choice of

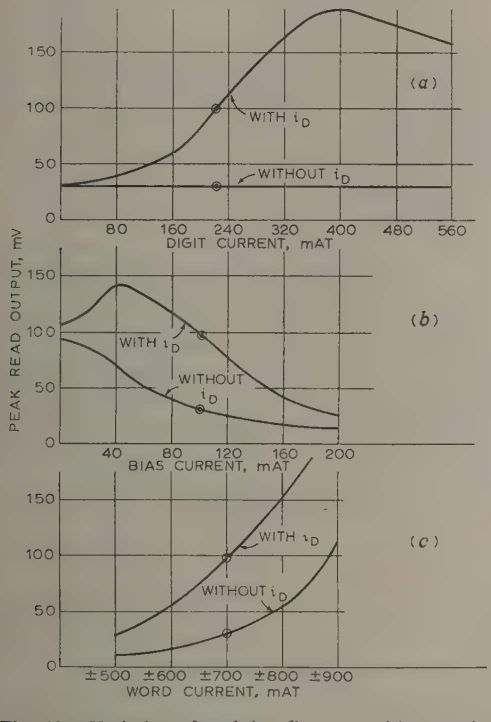


Fig. 12.—Variation of peak 'read' output with operating current.

(*a*) $i_w = \pm 700$ mA. $I_b = +100$ mA.
(*b*) $i_w = \pm 700$ mA. $i_D = -220$ mA.
(*c*) $I_b = +100$ mA. $i_D = -220$ mA.

digit current is restricted at low currents only by the minimum acceptable difference signal and at high currents by the decreasing effectiveness of the net reading current. However, i_D should not be much greater than $2I_b$ or the effect of repeated 'disturbing' digit currents on the quiescent operating points of the loop could cause deterioration of the stored information. Further tests have been carried out on the system with a nominal digit current of −220 mAT in order that non-selected words experience the maximum 'disturb' current likely to occur in practice.

(5.2.3) **Sensitivity to Operating Currents.**

The nominal word and digit currents are as follows:

$$i_w = \pm 700 \text{ mA.}$$
$$i_D = -220 \text{ mA.}$$
$$I_b = +100 \text{ mA.}$$

Fig. 11 shows the waveforms obtained under these conditions. The effect of variations in any one of the currents when the other two remain constant is shown in Fig. 12(*a*), (*b*) and (*c*). The difference signal from the core pair is fairly sensitive to small variations, and it is a disadvantage of this mode of operation that accurate definition of the operating currents is required.

(5.2.4) **Further Tests.**

The voltage-output waveforms of Fig. 11 indicate that the recovery of the cores to a stable position on the operating loops is not achieved until approximately 0·25 microsec after the 'write' current has been removed. If the system is to be operated at a 0·6 microsec cycle time, there are two possibilities to consider:

(*a*) If a given word is being operated at a maximum permissible rate, 'read' periods will occur at 0·6 microsec intervals, and thus a core will be read as soon as the preceding 'write' current pulse is removed, i.e. while the core is in an unstable state.

(*b*) If a sequence of different words are being operated at the maximum rate, then a given core pair will be acted on by the word and digit currents, and then receive a number of 'disturbing' digit pulses. The first of these 'disturb' pulses occurs 0·6 microsec after the active digit pulse on that core pair, when the cores are very near to their stable states.

It is found that the stability of the system is satisfactory under each of these circumstances. In the first case the 'read' current is more than sufficient to change the available flux, and in the second the restoring action of the bias is adequate. Changes in the 'read' output signals from the cores do occur under these circumstances, as would be expected, but they do not exceed 20% and can be considerably reduced by increasing the cycle time to 0·7 microsec.

Further increases in speed can be obtained by reducing the duration of the operating currents and tolerating a smaller flux change. However, the flux change could be maintained provided that the amplitudes of the operating currents were correspondingly increased. This has not been investigated in detail, since transistors capable of driving even a small number of words significantly faster than the cycle time achieved are not readily available. Operation of the system for tests described in Section 4.1.1 also proved satisfactory.

(5.3) Use of Cores with a Low Coercive Force

The mode of operation described in Section 5.2 has been tried with FX2140 cores, which have a very low coercive current (of the order of 90 mAT) to determine whether the operating currents would be reduced. Curves similar to those of Fig. 10 are obtained at rather lower word and bias currents. It was found, however, that the recovery of these cores after the 'write' current is removed was comparatively slow, of the order of 1 microsec, and the operating loop is therefore extremely sensitive to repeated digit pulses. These cores have, however, been used with success in a partial flux-switching system,[14] using

an unbalanced word current and no bias, in which the digit current is used to decrease the switching action of the 'write' current in one core of a core pair and to supplement its action in the other.

(6) CONCLUSIONS

Storage systems have been discussed which permit cycle times ranging from 0·6 to 2·0 microsec to be achieved. In a large store the limit is set, not by the characteristics of the core, but by the availability of transistors capable of driving the systems. At present, transistors are available which would permit the operation of a 1 024-word stack of core storage with a cycle time less than 2·0 microsec. To achieve significantly faster operation the size of the store must be reduced, and it is estimated that 128 words could be provided with a cycle time of about 0·6 microsec. It is expected that components will eventually become available which permit the operation of, say, 4 096 word stacks at a cycle time of about 1 microsec.

The most promising modes of operation considered are those which depend on partial flux switching. This enables the cores to be cycled at high repetition frequencies with relatively low power requirements, and heating effects in the cores are also minimized.

(7) ACKNOWLEDGMENTS

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(9) APPENDICES

(9.1) Memory and Switch Cores

(9.1.1) Memory Cores.

Four different types of 1·3 mm memory cores have been tested and their characteristics determined. The results relevant to the switching and driving requirements of these cores are summarized in Table 3.

Table 3

1·3 MM MEMORY CORES

Core type	I_f	T_s	H_c	K_r
	mA	microsec	oersteds	oersted-microsec
F764/S4	500	1·2	1·48	0·64
Development core, GC1	800	0·62	2·34	0·50
Development core, GC2	900	0·46	2·61	0·40
Development core, IB1	950	0·4	2·77	0·37

In this Table T_s is the switching time appropriate to the full current I_f, which is taken as $\frac{4}{3} I_c$, where I_c is the measured coercive current. The coercive force H_c is that defined by eqn. (1), and is calculated using the average radius.

The measured switching times of the three development cores, which have a higher coercive field than that of the F764/S4 core, indicate that cycle times of the required order could probably be obtained with a conventional half-current selection technique. The half-current would be of the order of 500 mA, and thus the driving requirements of such a system and those of the systems discussed in the paper would be similar.

(9.1.2) The F764/S4 Core.

The F764/S4 core is the only one currently available in quantity in this country, and attention has therefore been concentrated on this type. Fig. 13 shows the variation in switching time and peak output voltage with driving m.m.f. Also shown are the inverse switching time and the form of the flux change obtained under the test conditions. This plot of flux change is used as a calibration for the integrator used in many of the experiments.

The curves of Fig. 13, which were obtained by biasing the core to 500 mAT and driving against this bias with a pulse of variable amplitude, are the results of measurements taken on 50 cores, and they may be regarded as representing the average core.

(9.1.3) Partial Flux-Switching Measurements.

Fig. 14 shows the effect of switching currents which do not carry sufficient charge to permit complete reversal of the magnetic state of the core. The core is initially set to a standard

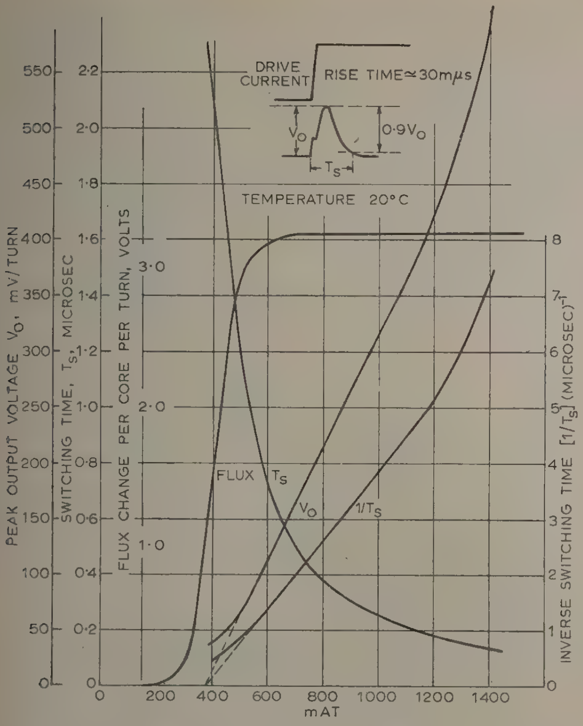

Fig. 13.—Switching properties of F764/S4 cores.

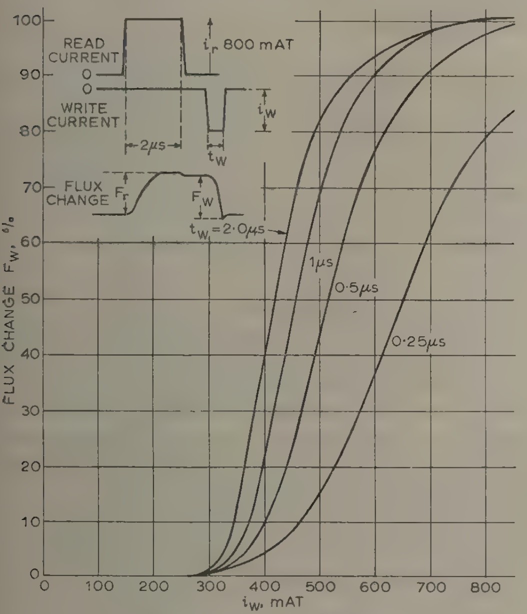

Fig. 14.—Partial flux switching.

saturated state by a current pulse i_r, of large amplitude and duration. It is then partially switched by a pulse i_w, the amplitude and duration of which are limited. The percentage flux change F_w, which occurs during the partial switching, is plotted as a function of the amplitude of i_w for various durations. For a given pulse duration there is an approximately linear relationship between F_w and current amplitude over a wide range. Over a limited range the flux changed by a given current is approximately proportional to its duration. For this type of operation F_r and F_w (Fig. 14) are approximately equal, which implies that the movement of the operating point back to the zero current axis of the minor loop [Fig. 16(*b*)] is associated with only a small flux change, which is virtually completed during the switching off of the current i_w.

An alternative method of achieving a partial flux switch is illustrated in Fig. 15. Here the currents i_r and i_w are maintained

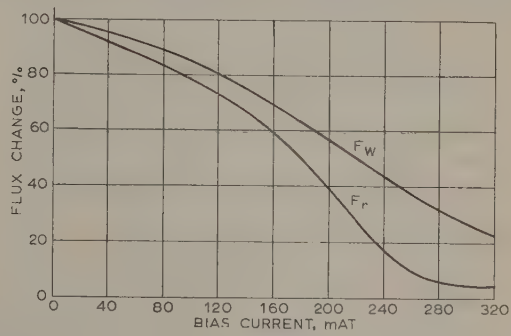

Fig. 15.—Partial flux switching with a bias current.

i_r = 800mA; 2·0microsec.
i_w = 800mA; 0·5microsec.
100% flux charge here corresponds to 97·5% in Fig. 14.

at constant amplitude and duration, and the flux changes appropriate to these currents are plotted as a function of a d.c. bias threading the core in such a direction as to diminish the switching action of i_w. Under these circumstances F_r and F_w are no longer nearly equal, in general $F_w > F_r$. The difference between F_w and F_r is made up during and after the switching off of i_w. A typical flux-change waveform and the postulated operating loop are shown in Fig. 16(*d*), which illustrates the type of flux-

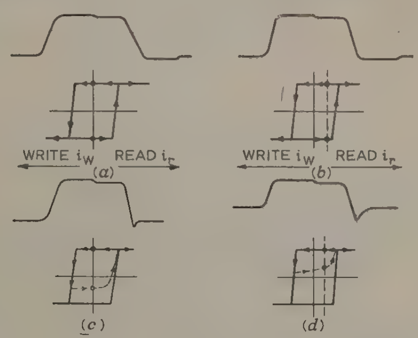

Fig. 16.—Flux-change waveforms and operating loops.

(*a*) Charge carried by i_r and i_w sufficiently large to allow saturation in each direction.
(*b*) Charge carried by i_w limited.
(*c*) System with bias; charge carried by i_r and i_w sufficiently large to allow saturation in each direction.
(*d*) System with bias; charge carried by i_w limited.

change waveform obtained under various conditions. There is an apparent anomaly between the curves of Fig. 14 and Fig. 15 in that the values obtained for F_w at a given net switching current (i.e. i_w in the first case, and i_w minus the bias current in the second) are not equal. This effect may be explained by noting that it is current in excess of the knee of the operating loop which is effective in changing flux. For the curves of Fig. 14 the knee has a constant value (taken here as 370 mAT, corre-

sponding to a 5% flux change) since the saturating current is constant. When a bias current is used, the saturating current, which is the sum of i_r and the bias current, increases as the bias increases, and so the effective 'knee' current also increases because of the increasing saturation of the core. A curve showing the variation of the 'knee' current, as defined above, with saturating current is shown in Fig. 17(a); points are included to give an indication of the experimental accuracy, which is limited because the variations measured are only a small percentage of the mean current. Using this curve a fair correlation is found between the curves of Figs. 14 and 15. This is illustrated in Fig. 17(b).

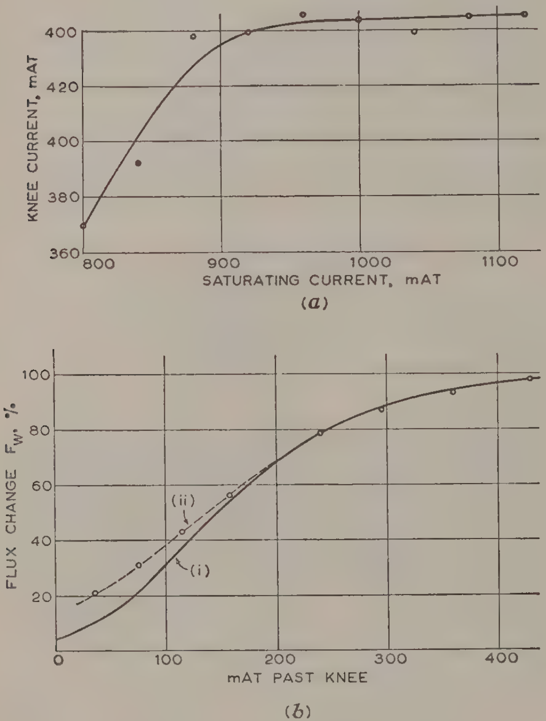

Fig. 17.—Correlation of Figs. 14 and 15.

(a) Variation of 'knee' current with saturating current.

(b) Variation of flux change with m.m.f. past knee.
(i) Saturating current = 800 mA, 2 microsec.
Knee at 370 mA.
'Write' current duration = 0·5 microsec.
(ii) Saturating current = 800 mA, 2 microsec + bias current.
Knee. See Fig. 17(a).
'Write' current = 800 mA, 0·5 microsec − bias current.

(9.1.4) **Switch Cores.**

The properties of the switch cores F002/S4 are of particular interest in relation to the properties of the F764/S4 core. The cores tested have the following dimensions:

	F002/S4 (L)	F002/S4 (S)
Outside diameter ..	2·49 mm	2·64 mm
Inside diameter ..	1·4 mm	1·5 mm
L_n	4·65 mm	2·59 mm
A	2·53 mm²	1·475 mm²
r_{av}	0·972 5 mm	1·035 mm
Volume	15·3 mm³	9·62 mm³
Outside dia. / Inside dia. ..	1·78 mm	1·76 mm

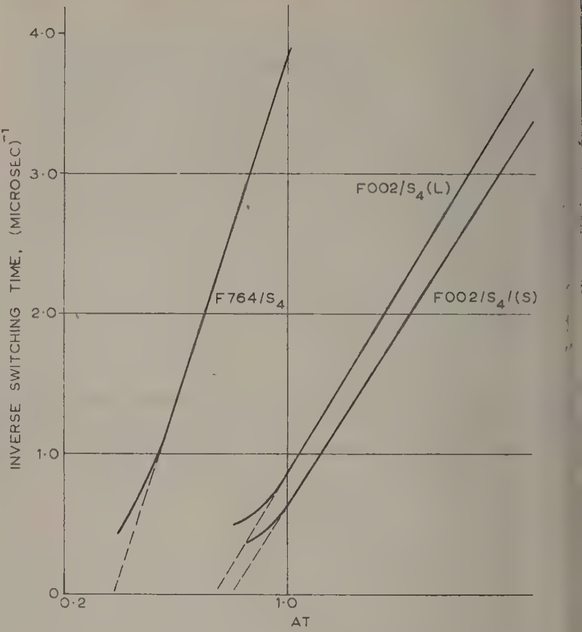

Fig. 18.—Switching times of cores in S4 material.

	I_c	K_s	H_c	K'_s
	mA	mA-µs	oersteds	oersted-µs
F764/S4 ..	380	160	1·48	0·625
F002/S4(L) ..	740	304	1·52	0·625
F002/S4(S) ..	800	320	1·54	0·62

The results of an experiment designed to yield values of th coercive force and switching coefficient are plotted in Fig. 18 The cores were biased to 1·95 oersteds and driven against th bias with a pulse of variable amplitude. (The bias field o 1·95 oersteds corresponds to 500 mAT for the F764/S4 core. The switching properties of the cores F002/S4 are virtuall identical with those of the F764/S4 core, the maximum deviatio being about 5%. This is an expected result since all the core are fabricated in the same material.

The relative flux density of the cores when they are cycle around similar hysteresis loops has been determined. Th experiment shows that at corresponding fields the flux densit of both the F002/S4 cores is greater than that of the F764/S core, and that the full saturation limits of the cores are no reached at a field of 1·95 oersteds (corresponding to the fu current of the F764/S4 core). The reason for the increased flu density of the F002/S4 cores is the increased density (g/cm³ of the S4 material when fabricated in this form. The densit of the material in the three types of core is given below, wher it can be seen that there is a correlation, within practical limit of measurement, between percentage increase in density an percentage increase in flux density:

	Density	Increase in density	Increase in flux density at 1·95 oersteds
	g/cm³	%	%
F764/S4 ..	3·83	—	—
F002/S4 (L) ..	4·06	6	7
F002/S4 (2) ..	4·7	23	26

).1.5) Effect of Temperature on Magnetic Properties.

As the temperature of the cores is increased it is found that ıe switching time for a given drive current and the remanent ux density of the material each decrease. The reduction in vitching time with temperature, which may be attributed to a ecreasing coercive force, tends to cause an increase in the eak 'read' output voltage as the temperature increases. The ecrease in remanent flux density has the opposite tendency, ut since the decrease in switching time is rather faster than ıe decrease in flux density, it is found that the peak output oltage increases with temperature over a wide range.

Fig. 19(*a*) shows the variation of the switching time, flux

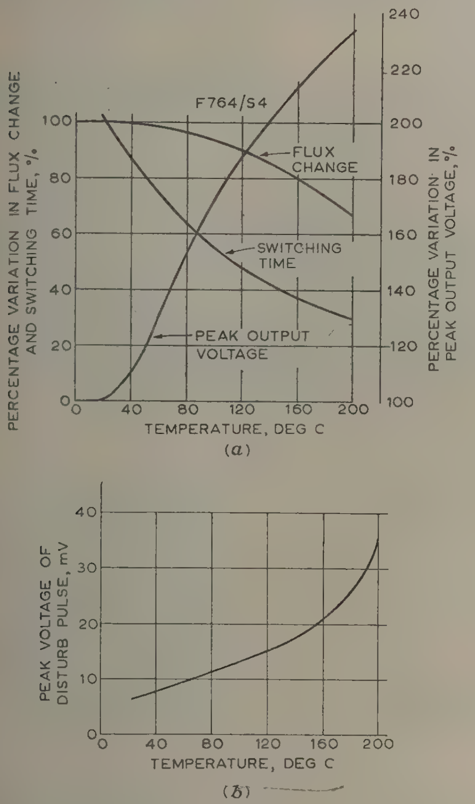

Fig. 19.—Variation of magnetic properties with temperature.
(*a*) Peak output voltage, flux change and switching time.
Switching current = 500 mA. Rise time ≃ 0·1 microsec.
(*b*) Peak 'disturb 1' output voltage.
'Disturb' current = + 250 mA. Rise time ≃ 0·1 microsec.

ıange and peak output voltage of the F764/S4 core over a) : 1 range of temperature, plotted as a percentage of the ılues measured at room temperature (20° C). In Fig. 19(*b*) ıe variation with temperature of the amplitude of the isturbed 1' pulse is shown. This curve indicates that the slope the relatively flat portions of the *B*/*H* loop increases with mperature. At the higher temperatures (160° C and above) ıe increase in output amplitude becomes more rapid, and reversible flux changes occur indicating that the 'disturb' ırrent is moving the operating point beyond the knee of the op.

(9.2) Theory of Flux Switching and Design of a Switch Core

.2.1) Flux Switching.

In Fig. 20(*a*) a switch core S is shown linked to *n* memory ɔres. R_s represents the 'wire resistance' of the secondary plus

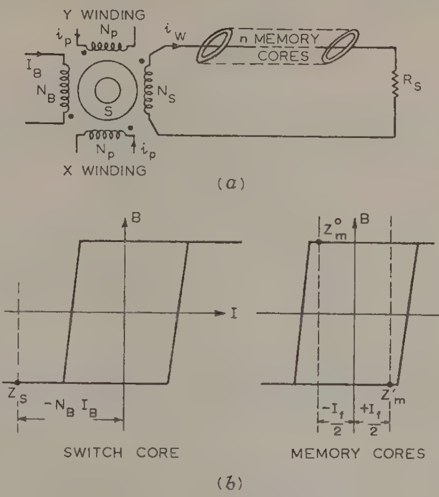

Fig. 20.—Flux switching.

any extra resistance included for measurement or other purpose. Initially it is assumed that the switch core is biased to some point Z_s on the *B*/*H* loop, and the memory cores are biased to the 'half-current' point [Fig. 20(*b*)]. These conditions are in accordance with bias system A discussed in Section 4.

When the switch core in an element is a matrix the primary m.m.f. will be supplied by two windings X and Y, and when both windings are operative the total m.m.f. supplied is sufficient to overcome the bias and reverse the flux in the switch and memory cores. On removal of the primary drives the switch core reverts to its original state under the influence of the bias, whilst the switching in the memory cores is assisted by a digit current which effectively reverses the d.c. bias on these cores.

The basis for the analysis is the equation

$$N_s\frac{d\Phi_s}{dt} \times 10^{-8} = n\frac{d\Phi_m}{dt} \times 10^{-8} + i_wR_s + L_s\frac{di_w}{dt} \quad . \quad (3)$$

which may be rewritten as

$$N_sA_s\frac{dB_s}{dt} \times 10^{-8} = nA_m\frac{dB_m}{dt} \times 10^{-8} + i_wR_s + L_s\frac{di_w}{dt} \quad . \quad (4)$$

where A_s and A_m are the cross-sectional areas of the switch and memory cores and L_s is the inductance of the word wire loop.

Since the behaviour of the cores when switched at high speed is non-inductive, further consideration of eqn. (4) in its present form is difficult because of the terms in R_s and L_s. Provided that R_s and L_s are sufficiently small an approximation to eqn. (4) is

$$N_sA_s\frac{dB_s}{dt} = nA_m\frac{dB_m}{dt} \quad . \quad . \quad . \quad . \quad . \quad (5)$$

Integrating this equation,

$$N_sA_s\Delta B_s = nA_m\Delta B_m \quad . \quad . \quad . \quad . \quad (6)$$

If a choice of switch core is allowed the quantity N_sA_s is a design variable, and assuming that the cores are of the same material $\Delta B_s = \Delta B_m$. In this case, and when a complete flux change in the memory cores is required,

$$N_sA_s = nA_m \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

For a complete flux change eqn. (5) then becomes

$$\frac{dB_s}{dt} = \frac{dB_m}{dt} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5a)$$

This equation enables the driving requirements of the system to be estimated, since, if the switching rates of the cores are equal, the switching times will be the same.

Therefore $$H_s = H_m \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (8)$$

where H_s is the net field which has to be applied to the switch core in order that the memory cores shall switch in a time appropriate to the field H_m. In eqn. (8) both H_m and H_s are measured from the origin.

(9.2.2) **Design of a Switch Core.**

As an example, the design of a switch core to drive fifty-two F764/S4 memory cores is considered. The design equation is

$$N_sA_s = nA_m \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

and the following practical points must be noted.

(*a*) There will be n unswitched cores on the secondary loop which will absorb a small fraction of the available flux change from the switch core, and R_s and L_s may absorb an appreciable fraction of the available flux change. This is not taken into account in the design considered here, although the unswitched cores are allowed for by increasing the value of A_m by 5%.

(*b*) The 'squareness', i.e. the slope of the steep part, of the B/H loop of the switch core should be similar to that of the memory cores; here the ratio of outside to inside diameter is made the same as that of the memory cores.

(*c*) The average radius of the switch core must be kept comparable with that of the memory cores, or the m.m.f. required to produce a given field will be excessive, but the average radius must be large enough to allow sufficient winding space.

(*d*) It is possible to fabricate a switch core with a ratio of length to outside diameter of about 2 : 1.

The dimensions of the F764/S4 core are

Outside diameter $= 1\cdot3$ mm $\quad A_m = 0\cdot1045\,\text{mm}^2$
Inside diameter $= 0\cdot75$ mm $\quad r_m = 0\cdot5125$ mm

$$L_n = 0\cdot38\,\text{mm} \qquad \frac{\text{Outside diameter}}{\text{Inside diameter}} = 1\cdot73$$

Then choosing $N_s = 2$

$A_s = n1\cdot05A_m$ from eqn. (7)

Therefore $A_s \simeq 2\cdot85\,\text{mm}^2$

The average radius is given by

$$r_s = \frac{\text{Inside diameter}}{2} + \frac{\text{Outside diameter–Inside diameter}}{4}$$

and $$\frac{\text{Outside diameter}}{\text{Inside diameter}} = 1\cdot73$$

Choosing $r_s = 2r_m$

Outside diameter $= 2\cdot6$ mm
Inside diameter $= 1\cdot5$ mm

But $$A_s = \frac{\text{(Outside–inside) diameter}}{2} L_n$$

Therefore $L_n = 5\cdot2$ mm.

The dimensions of the required core are therefore

Outside diameter $= 2\cdot6$ mm
Inside diameter $= 1\cdot5$ mm
$L_n = 5\cdot2$ mm

(9.2.3) **Primary Ampere-Turn Requirements of Bias System A.**

The switch core F002/S4 (L) has dimensions which appro
mate to those specified, although the cross-sectional area
rather smaller than this design requires. This is not a seve
disadvantage, for two reasons:

(*a*) The flux density of the S4 material when fabricated in t
form is rather greater than measurements on the memory cores
S4 material indicate.

(*b*) The switch cores operate on a larger loop than the memo
cores, which do not attain full saturation even when the operati
loop is that recommended for normal 'half-current' selection.

Tests on this switch core indicate that the switching coefficie
and coercive force are virtually identical with those of t
F764/S4 core (see Appendix 1) and it is therefore permissible
use eqn. (8) to estimate the driving requirements.

For Bias System A

Memory core bias, $i_w = \pm\, 500$ mA in one turn
$I_b = +\, 250$ mA in one turn
$i_D = -\, 500$ mA in one turn

and $N_s = 2$

Since the 'read' and 'write' currents are of equal magnitude,

$$N_pi_p = 2N_Bi_B \quad . \quad . \quad . \quad . \quad . \quad .$$

where N_pi_p is the total applied primary ampere-turns.

Eqn. (8) then becomes, for the 'read' period,

$$\frac{\frac{N_pi_p}{2} - N_si_w}{r_s} = \frac{i_w + I_b}{r_m} \quad . \quad . \quad . \quad . \quad ($$

and a similar equation holds for the 'write' period. Substituti
values in eqn. (10),

$$N_pi_p = 2\cdot43\,\text{AT}$$

Thus the required bias on the switch core is $2\cdot43$ AT and ea
primary drive is of the same magnitude as the bias.

[The discussion on the above paper will be found on page 605.]

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HIGH-SPEED LIGHT OUTPUT SIGNALS FROM ELECTROLUMINESCENT STORAGE SYSTEMS

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SUMMARY

The possibility that a matrix of electroluminescent cells could be used for storage of information in digital form has long been realized. The problems involved in reading information from a simple type of permanent store are discussed. It has been shown that the read-out rate is limited by the afterglow of the phosphor, and, for a particular (ZnS, Cu, Cl) phosphor, it is 25–30 microsec per digit. A theoretical calculation of the discrimination ratio, i.e. the ratio of the output signal from a '1' to the interference signal obtained when a '0' is selected, shows that it should be possible to operate matrices with as many as 64^2 cells. This conclusion is not borne out by measurements made on experimental panels; these suggest that 32^2 is about the maximum. Results obtained in the paper are regarded as preliminary since arbitrary choices have had to be made concerning some aspects of the operation of the matrices. There is little doubt that considerable improvement in performance is possible.

LIST OF SYMBOLS

a = Length of the side of a cell in an electroluminescent matrix.
c = Separation of adjacent cells in a matrix.
d = Distance from the centre of a matrix to a corner cell.
r = Perpendicular distance from the centre of a matrix to the centre of the photocathode.
p = Radius of the photocathode.
t = Thickness of the glass substrate on which the electroluminescent cells are mounted.
x = Parameter used to define a method of selecting the cells in a matrix.
V = Amplitude of the excitation voltage pulse applied to a cell, volts.
V_c = Output pulse which results from integration of the photocurrent.
B = 'Brightness' or integrated light output of an electroluminescent cell excited by a voltage V.
T = Time for the amplitude of an emitted light pulse to fall to one tenth of its maximum, microsec.
T' = Period by which the integration pulse must be delayed for V_c to be one tenth of its maximum value.
V_w = Value of V_c when a '1' is selected in the matrix.
V_u = Value of V_c when a '0' is selected in the matrix.
θ = Angle subtended at the axis of the photocathode by a corner cell on the matrix.
θ_c = Critical angle of incidence for glass to air.
σ = Standard deviation.

(1) INTRODUCTION

Several papers[1, 2, 3, 4] have appeared during the last two years describing electroluminescent panels. These have been concerned with visual displays, but there has, for a long time, been a realization that it may be possible to use similar panels for digit storage. In particular, a permanent or semi-permanent type of store for an electronic digital computer may be visualized. In its simplest form this could consist of a matrix of electroluminescent cells masked by a punched card or a photographic negative. The information in such a store could be altered only by changing the mask. It could be read out by selecting the cells one at a time by means of appropriate voltage pulses, and detecting the resultant light flashes with a photomultiplier. Photographic masks could be made with high accuracy by exposure to the electroluminescent panel on which the required pattern was being displayed.

A more flexible store may be made if the electroluminescent phosphor is replaced by a matrix of two-state electroluminescent/photoconductive switches. These two-state switches consist of an electroluminescent and a photoconductive cell electrically in series, with the photoconductive-cell receiving part of the light emitted by the electroluminescent cell. The properties of switches of this type (optrons) have been discussed in several papers.[5, 6, 7] Information stored in such a matrix can be read rapidly (a few microseconds per digit). Writing new information into the store involves selection of each element in turn. As the time taken to change from the on state ('1') to the off state ('0') is about 0·01 sec, writing is essentially a slow process, but switching times are being improved as better photoconductive techniques become available.

Possible advantages of using an electroluminescent matrix or a matrix of photoconductive-electroluminescent elements include the following:

(a) Each element may be made very small so that relatively large numbers of digits can be stored in a small area.
(b) The method of construction is suitable for the production of cells which are very uniform, at a relatively low cost.
(c) The read-out may be very rapid, and is obtained in the form of a pulse of light. This can be expected to simplify pick-up problems.

If such a store—or any other store utilizing the emission from electroluminescent cells—is to be used in a digital computer, it is necessary to know the limitations which are inherent in the nature of the electroluminescent light pulse. It is the purpose of the paper to discuss the response of electroluminescent cells to pulses of voltage, and to estimate the read-out speed and the number of cells which could be tolerated in an electroluminescent matrix.

(2) ELECTROLUMINESCENCE

Electroluminescence, the sustained emission of light from a phosphor subjected to an alternating electric field, has been the subject of many papers, including References 8–12. A typical electroluminescent cell is shown in Fig. 1. When an alternating voltage is applied to the electrodes, light is emitted, and it emerges through the transparent bismuth-oxide/gold

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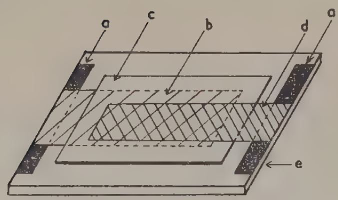

Fig. 1.—Electroluminescent cell.

a. Silver contact.
b. Transparent gold on bismuth oxide.
c. Phosphor layer.
d. Copper.
e. Glass.

layer. Contact is made to the electrodes by soldering directly to silver strips baked into the glass substrate. The phosphor layer is, in fact, two layers. The first is a suspension of zinc-sulphide phosphor in nitrocellulose, and on top of this is a layer consisting of barium titanate in a resin binder.

To make a matrix of cells, the single electrodes are replaced by narrow conducting strips, as shown in Fig. 2. The strips are produced by a photo-etching process, and each has a silver contact 'flag'.

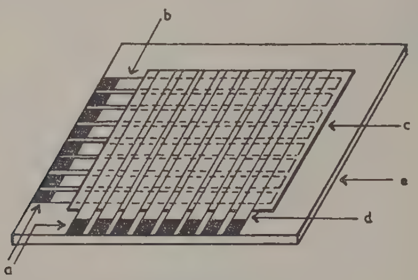

Fig. 2.—Electroluminescent matrix.

a. Silver contact 'flag'.
b. Transparent gold bars.
c. Phosphor layer.
d. Copper bars.
e. Glass.

Although the response of cells such as these to sine waves and square waves is well known, there has been very little published work on the response to narrow voltage pulses. It was anticipated that pulses of less than 10 microsec duration would be used to excite the individual cells in a matrix, and so the properties of the light pulses emitted by electroluminescent cells under these conditions have been studied.

A cell excited at regular intervals by a narrow pulse emits a series of double flashes of light. There is a flash at each edge of the excitation voltage pulse. If the light is detected by a photomultiplier and the photocurrent is amplified and displayed on a cathode-ray oscillograph, waveforms similar to those of Fig. 3 are obtained.

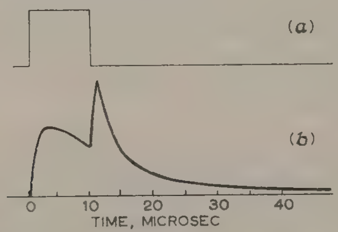

Fig. 3.—Light waveform for pulse excitation.

(*a*) Excitation voltage pulse.
(*b*) Light emission.

The properties of the light pulses of most interest in study are as follows:

(*a*) Individual pulses may vary widely. The varia decreases as the brightness of the emitted light increases.

(*b*) Pulses may rise to a maximum in as little as 0·1 micr although the onset of the pulse does not occur immediatel the start of a voltage edge.

(*c*) When the cell is excited by a 5 microsec pulse, the sec flash of each pair decays to one third of its maximum bright in 2–3 microsec, and to one tenth in about 18–25 microsec room temperature). The decay time is dependent both pulse-repetition frequency and temperature, and, in genera increases as the pulse width increases.

(*d*) Increasing the amplitude of the excitation voltage p has only a slight effect on the rise and decay times of the l pulses.

(*e*) As the amplitude of the voltage pulse is varied, the va tion in the maximum amplitude of the light pulse can expressed approximately by a relationship of the form

$$B = A \exp(-b/\sqrt{V}) \quad . \quad . \quad . \quad . \quad .$$

where A and b are parameters which depend on the phosp the cell construction and the experimental conditions, and the amplitude of the excitation voltage pulse. It has long known[9] that a similar formula describes accurately the dep dence of the 'brightness' or 'integrated light output' of a excited by a wide variety of waveforms.

(3) READING OUT FROM AN ELECTROLUMINESCEN PANEL

The store which will be considered is the simplest poss namely a matrix of $n \times n$ electroluminescent cells with a m which has holes to indicate '1's and is opaque where '0's stored. This is representative of any store in which the infor tion is read by detecting the light pulses from an elec luminescent cell.

(3.1) Size of the Matrix

It will be assumed that it is desirable to have as many as possible in a matrix, and that each cell should be as sma possible. In order to ensure that a large portion of the emi light reaches the photomultiplier cathode the matrix shoul as close to the cathode as possible. In a practical system of two situations arises: either the area of the matrix is than that of the cathode, in which case it can be extrem close to it, or the matrix is bigger, when it must be far eno away for there to be very little difference in the amount of l reaching the cathode from cells situated at different posit in the matrix.

Suppose initially that an $n \times n$ matrix is bigger than photocathode, and that each of its cells is square, of sid and separated from its neighbours by a distance c. Let perpendicular distance from matrix to cathode be r, as sh in Fig. 4. It is clear that the cathode will receive more l from a cell on the common axis than from an identical ce

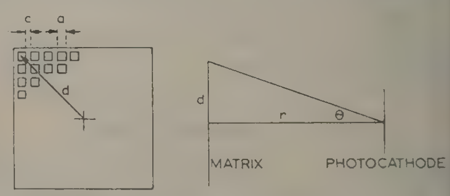

Fig. 4.—Spacing of cells and position of photomultiplier.

corner of the matrix. If the cells are regarded as radiators ich obey Lambert's law, and if $r \gg d$, the distance from the tre to the corner of the matrix, the photomultiplier will eive $\cos^4 \theta$ times as much light from the corner cell as from axial cell, where

$$\tan \theta = \frac{d}{r} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

$$d = \sqrt{2}(n/2 - 1)(a + c) \quad . \quad . \quad . \quad . \quad . \quad (3)$$

can decide that $\cos^4 \theta$ must be at least as great as some itrary fraction. For example, if $\cos^4 \theta \geqslant 0{\cdot}7$, then $\theta \leqslant 24°$, from eqns. (2) and (3),

$$r \geqslant 3{\cdot}2(n/2 - 1)(a + c) \quad . \quad . \quad . \quad . \quad (4)$$

is convenient, when making the negative for the photo- ing process, to use a constant ratio of a to c, so that reduced nlarged copies can easily be made. In the matrices used in investigation $a = 2c$, and thus, for a 32×32 matrix,

$$r = 72a \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

for a 16×16 matrix,

$$r = 33{\cdot}6a \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (6)$$

$/a$ remains constant, in accordance with eqn. (5) or (6), e is no theoretical restriction on the size of the individual . This is because light reaching the photocathode from an l cell will be proportional to a^2/r^2, and will be constant. varies while r/a remains constant, θ is constant. When a duced so that r is not very much greater than p (the radius he photocathode) it can be shown that the cathode receives e light from a corner cell than would be predicted by the θ formula. If a is reduced until $d < p$, the restriction on appears and it can be made as small as is practicable.

is obvious that the size of individual cells is not limited by iderations of the uniformity of the light output. The final ation is likely to be the difficulty experienced in making cells ller than about 0·25 mm square. There is also a limit to sensitivity of the detecting equipment, and the smaller the , the greater must be the amplitude of the excitation voltage e. As shown in the next Section it is advantageous to use nall a pulse as possible. Matrices used for measurements ribed in a later Section were made with 1 mm-square cells, matrices with cells of 0·3 mm square have been made up other purposes.[1] If the photomultiplier with a cathode of m diameter is used, a 64×64 matrix of these 0·3 mm would be smaller than the cathode.

ius the number of cells in a matrix will be determined by the imination which can be obtained, and when this is known, size of the cell can be chosen so that $d < p$, if this is ticable. If it is not, eqn. (4) must be used to determine r.

iis conclusion is justified if the mask is in contact with the for instance if the back electrode were transparent. Using onstruction described in Section 2, the mask is necessarily rated from the cell by the thickness of the glass plate. If d and the matrix is very close to the photocathode, then, iown in Fig. 5, light from a square cell of side a appears e cathode to be emerging from the glass surface from an $(a + 2t \tan \theta_c)$ square, where t is the thickness of the glass θ_c is the critical angle of incidence. When a mask such as shown in Fig. 6 is being used, the light from a selected '0' not be able to escape through an adjacent '1'.

erefore

$$t \tan \theta_c < c \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (7)$$

r a 1 mm-square cell eqn. (7) implies that the glass must not icker than about 0·33 mm. When glass thicker than that

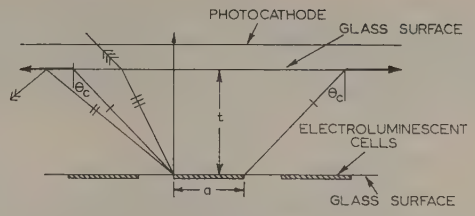

Fig. 5.—Effect of refraction in the glass substrate.

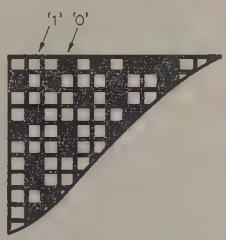

Fig. 6.—Part of a typical mask.

specified by eqn. (7) has to be used, the matrix must be withdrawn until the angle of refraction of any rays which emerge from a '1' adjacent to a selected '0' is great enough to make them miss the photocathode altogether. This is illustrated by the ray ABC in Fig. 7, and the critical distance r_0 can be calculated from the geometry of the figure.

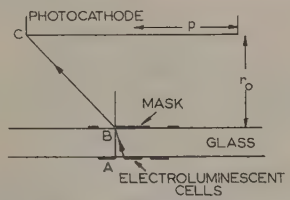

Fig. 7.—Minimum distance from photocathode.

(4) SELECTION OF A CELL AND DISCRIMINATION

The 'crossed conductor' method of construction described in Section 2 enables one to select a particular cell (i, j) by applying an electric field between the ith horizontal conductor and the jth vertical conductor. Some or all of the other $(n^2 - 1)$ cells will experience a smaller electric field. The amplitude of this, and hence the amount of light emitted by these unselected cells, will depend on the way the unselected row and column conductors are connected. It must be possible in the extreme case to discriminate between the light pulse obtained when a '1' is selected from a matrix which has $(n^2 - 1)$ '0's and the pulse which occurs when a '0' is selected in a matrix in which all the non-selected cells are '1's.

The 'discimination ratio' was discussed in Reference 1, and it can be shown that the selection scheme suggested therein, namely to apply $+v/2$ and $-v/2$, respectively, to the selected row and column, and $-v/6$, $+v/6$, respectively, to all the non-selected rows and columns, gives the greatest possible discrimination when a number of selected cells are being photographed on the same film.

Using a photomultiplier to read out from a matrix is quite a different matter, and the method of selection which gives the best discrimination can be found as follows.

(4.1) Discrimination Ratio

Let the excitation voltage pulses, as shown in Fig. 8, be applied to the conductors of the matrix. Thus the selected-row conductor receives a pulse which rises from 0 to $+v/2$ with respect

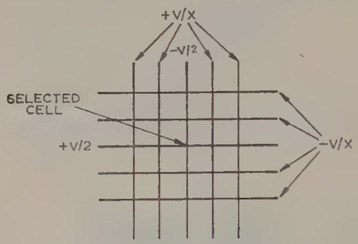

Fig. 8.—Method of selection.

to earth, and the pulse on the selected column goes from 0 to $-v/2$ with respect to earth. All the non-selected row conductors are connected together and receive a pulse which goes from 0 to $-v/x$, and the non-selected columns one which goes from 0 to $+v/x$. Therefore the selected cell has a pulse of amplitude v. The $2(n-1)$ cells along the selected row and column have pulses of $(v/2 - v/x)$. The remaining $(n-1)^2$ cells have pulses of $(2v/x)$. The limiting cases are when

(i) $x \to \infty$, i.e. the non-selected conductors are all earthed. There are then $2(n-1)$ unselected cells with pulses of amplitude $v/2$, and $(n-1)^2$ have no voltage across them.

(ii) $x \to 6$, when all (n^2-1) unselected cells have pulses of amplitude $v/3$.

Each cell will emit a light pulse which has a maximum intensity given by eqn. (1), i.e.

$$B = A \exp(-b/\sqrt{v})$$

If V_w is the 'wanted' signal, i.e. the output from the photomultiplier resulting from the light emitted by a selected '1',

$$V_w = A' \exp(-b/\sqrt{v}) \quad . \quad . \quad . \quad . \quad (8)$$

The 'unwanted' signal, V_u, is the sum of outputs which would result from all (n^2-1) unselected '1's when a '0' is selected, i.e.

$$V_u = 2(n-1)A' \exp\left(-\frac{b}{\sqrt{v}}\sqrt{\frac{2x}{x-2}}\right) + (n-1)^2 A' \exp\left(-\frac{b}{\sqrt{v}}\sqrt{\frac{x}{2}}\right) \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (9)$$

$$\frac{V_w}{V_u} = \frac{1}{2(n-1)\exp\left[-\frac{b}{\sqrt{v}}\left(\sqrt{\frac{2x}{x-2}}-1\right)\right] + (n-1)^2 \exp\left[-\frac{b}{\sqrt{v}}\left(\sqrt{\frac{x}{2}}-1\right)\right]} \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$

Using eqn. (10) curves of V_w/V_u as a function of x have been drawn for $b/\sqrt{v}$ varying from 8 to 17, with $n = 16$, 32 and 64. Some of these curves are reproduced in Fig. 9. They show, for example, that V_w/V_u will be greater than 10 for $n = 32$, when $b/\sqrt{v} = 11$.

The value of x which gives the maximum discrimination (maximum value of V_w/V_u) has been obtained independently, using the Newton–Raphson iterative process, and the results are given in Table 1.

Experimentally observed values of $b/\sqrt{v}$ ranged from about 10 to 12, and it can be seen that in this range, for any value of n between 16 and 64 which gives a usable value of V_w/V_u, the optimum value of x is close to 8. Indeed, it is very likely that in any experimentally realizable conditions the best value of x is likely to be between 7 and 8·5.

Using a blue/green-emitting phosphor (ZnS : Cu, Cl) with cells of the type described in Section 2, values of b ranging from 130 to 160 have been observed, and, depending on the experimental

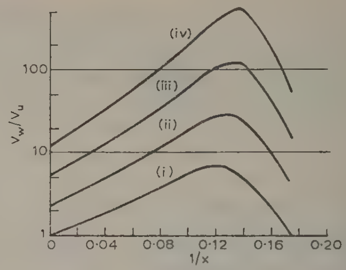

Fig. 9.—Discrimination ratio for a 32 × 32 matrix.

(i) $b/\sqrt{v} = 10$.
(ii) $b/\sqrt{v} = 12$.
(iii) $b/\sqrt{v} = 14$.
(iv) $b/\sqrt{v} = 16$.

Table 1

Optimum Value of x

Optimum value of x for specified $b/\sqrt{v}$										
$b/\sqrt{v}$ =	8	9	10	11	12	13	14	15	16	1
$n = 16$	8·38	8·01	7·74	7·53	7·37	7·24	7·13	7·03	6·96	6·
$n = 32$	9·14	8·65	8·28	8·00	7·78	7·61	7·46	7·34	7·24	7·
$n = 64$	9·94	9·30	8·84	8·49	8·21	7·99	7·81	7·65	7·53	7·

conditions, an adequate value of V_w can be obtained fr pulses with $V = 120$–136 volts, when matrices with 1 mm-squ cells are used. Thus it is quite possible to obtain a value $b/\sqrt{v}$ of greater than 12, and so, on the basis of these calculati it seems possible that 64 × 64 matrices could be used.

(5) READ-OUT SPEED

As Fig. 3 shows, the light pulse initiated by the edge c square pulse of voltage rises very rapidly. In a typical case light pulse may start rising before the excitation voltage edg completed, and it may reach its maximum in 0·2 microsec less. It then falls relatively slowly, and, depending on a num of variables, may take from 15 to 50 microsec to decay to tenth of its maximum. It is this 'afterglow' which limits rate at which digits may be read from an electroluminesc matrix. When a number of cells are selected in succession, photomultiplier detects light emitted from the selected cell, also the light from any others which have been selected viously. The digit period must therefore be made long eno to ensure that the residual emission from all previously selec cells is much less than the output of a selected '1'. In o words, it must be possible to distinguish between a '1' wh follows a series of '0's and a '0' which follows a succession of '

(5.1) Integrating the Light Signal

Fig. 3 shows the waveform of the voltage pulse whic developed across a 1-kilohm load in the anode of a ph multiplier when the cathode detects a pulse of light from

troluminescent cell. This waveform is not suitable for rating the pulse circuits in a digital computer.

ntegrating the light signal has the effect of improving the rimination—as, of course, has any adjustment which reduces amplitude of the excitation voltage pulse necessary to produce inimum usable output. This is because the factor $b/\sqrt{v}$ in . (10) is increased when v is reduced. The choice of 5 microsec the integration period was made as a compromise between need to obtain a high discrimination ratio and the desire to as a short a digit period as possible. To perform this ction, the diode bridge circuit shown in Fig. 10 was used in

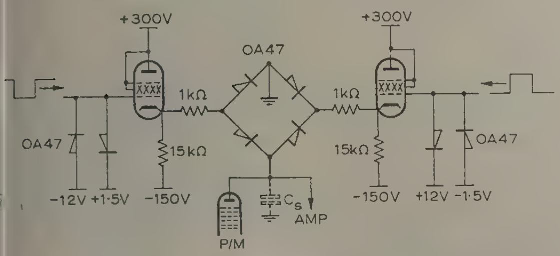

Fig. 10.—Integration of the photocurrent.

anode of the photomultiplier. During the integration od, two identical strobe pulses cause the diodes to be non-ducting, and the stray capacitance, C_s, is charged by the tocurrent to a voltage V_c. During the rest of the period diodes conduct and there is effectively a resistance of about ohms across C_s, which discharges very rapidly to a much er voltage. The strobe pulses are synchronized with the age pulses used to excite the electroluminescent cell, and the ipment was designed so that they can be varied in width, delayed by a variable time. It was found that V_c was test when the integration period started just at the end of the tation voltage pulse.

.2) Choice of Pulse Width and Pulse-Repetition Frequency

ne of the features of electroluminescence is the phenomenon uild-up.[1, 9, 12] If after a period of rest a cell is excited by a titive waveform, it is found that the light pulses gradually ease in amplitude over the first few cycles until an equilibrium eached. Similarly, if a cell is excited by irregularly spaced es of voltage, it is found that the light emitted at each pulse ends on the previous 'history'. For this reason, in the paper ntinuous sequential selection is considered, so that each cell lected once in every operating cycle.

he discriminations calculated in Section 4 are such that it ns likely that a 32 × 32 matrix could be operated success-, and so the repetition rate has been based on 1 024 digit ods. The width of the excitation voltage pulse influences n the amplitude of the light pulse and its decay time. A e width of 5 microsec was chosen as it was found that the put signal, V_c, doubled when the pulse width was increased n 1 to 5 microsec, and only increased by a further 40% n the pulse was increased from 5 to 20 microsec.

onsider the sequential selection of successive digits, each cating a '1'. It was decided that the excitation of these ts should be separated (by a time T') until the contribution V_c from a '1' which had been excited in the previous digit od was one tenth of the output to be expected from the cted '1'. The integrated photocurrent pulse V_c for a fixed gration period, of course, progressively decreases in ampli- as the start of the integration period is delayed after the excitation voltage pulse. As Table 2 shows, T' is less sensitive to variations in excitation pulse width than the light output signal, but is greater than T (the time taken for the light output signal to decay to one tenth of its maximum value). Using a

Table 2

Effect of Excitation Pulse Width

Pulse width	V_c	T	T'
microsec	volts	microsec	microsec
1	1·4	17	22
2	2·1	18	25
5	3·0	20	29
10	3·6	22	31
20	4·3	25	35

5 microsec pulse and integrating the photocurrent for 5 microsec results in T' being about 30 microsec, and thus a repetition frequency of 32 c/s was chosen.

(5.3) Improvement in Read-Out Speed

With the phosphor at present available, measurements made over a range of temperature from −80 to +120° C showed that the decay time depends in a complicated way on the temperature. In general, when a cell is being excited with narrow voltage pulses at a low repetition rate, the light pulse decays more rapidly at higher temperatures. Operating a panel at 100° C, for example, might lead to a reduction of 20% in the digit period. Symmetrical double pulses of voltage (Fig. 11) were

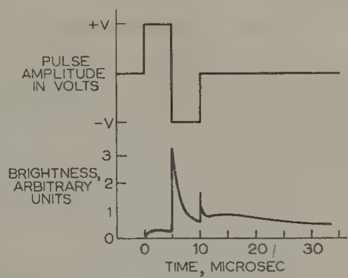

Fig. 11.—Brightness waveform for a double pulse.

also tried, but the decay time was greater than for single pulses except at temperatures greater than 100° C, when it approached the same value. It seems likely that a really significant improvement in read-out speed could be obtained only by the use of a phosphor which has inherently a shorter afterglow.

(6) EXPERIMENTAL OBSERVATIONS ON ELECTROLUMINESCENT MATRICES

(6.1) Variation of $\overline{V_c}$ as a Function of Pulse Amplitude

Measurements of $\overline{V_c}$, the average value of V_c, for a wide range of voltage-pulse amplitudes have been made on a number of small electroluminescent cells, on different cells selected from both 16 × 16 and 32 × 32 matrices, and on groups of cells from these matrices. Fig. 12 shows some typical curves of V_c plotted on a logarithmic scale, as a function of $1/\sqrt{v}$, where v is the amplitude of the 5 microsec exciting pulse. In all cases the integration period was 5 microsec, and the pulse-repetition frequency was 32 c/s. The curves approximate closely to

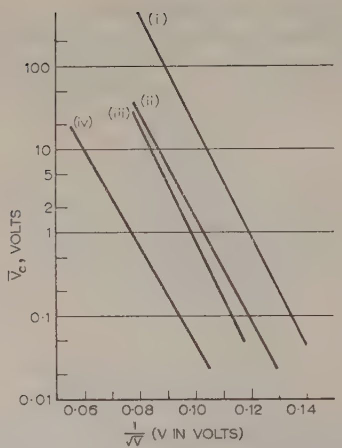

Fig. 12.—Integrated output as a function of the excitation pulse voltage.

(i) Group of cells (12), 2 × 2 mm at 15 cm.
(ii) Single cell 1 mm² at 1 cm.
(iii) Single cell 1 mm² at 5·2 cm.
(iv) Single cell 2 × 2 mm at 6 cm.

straight lines, showing that V_c is approximately related to v by an expression similar to eqn. (1), namely

$$\overline{V_c} \simeq A' \exp(-b/\sqrt{v})$$

The slopes of the lines in Fig. 12 give the appropriate values of b. For different cells b varied within the range 130–160.

(6.2) Statistical Distribution of V_c

In the calculations on discrimination it was implicitly assumed that V_w and V_u were, for any given operating conditions, constant from one pulse to another. But as described in Section 2, the light output from a succession of identical excitation pulses fluctuates in an apparently random manner. Since it is essential when reading out from a matrix that a '1' should always be distinguishable from a '0', it is necessary to know that the variation in V_w and V_u will not be too great. The variability of the light pulse from a single cell in a 32 × 32 matrix as a function of the amplitude of the voltage pulse has been studied by recording V_c on a cathode-ray oscillograph with a single-sweep time-base. This cell was continuously excited at a pulse-repetition frequency of 32 c/s, and the time-base was triggered at 1 sec intervals. 500 traces were measured at each of five voltages, and the results have been analysed. Table 3 gives the averages, the standard deviations, σ, and the ranges within which 95% of observations could be expected to be found ($\overline{V_c} \pm 2\sigma$).

Table 3

VARIABILITY OF V_c

V	$\overline{V_c}$	σ	$\overline{V_c} \pm 2\sigma$	$\sigma/\overline{V_c}$
volts	volt	volt	volt	
120	0·098	0·036	0·026–0·170	0·37
160	0·89	0·25	0·39–1·39	0·28
190	2·04	0·38	1·28–2·80	0·19
214	3·8	0·55	2·70–4·90	0·14
239	6·19	0·36	5·47–6·91	0·06

Table 3 shows that the standard deviation becomes relatively larger at the lower pulse amplitudes. In operation the individual cells just give an output which is reliably observable with measuring equipment used, at 120 volts excitation, and could operated in the range up to 240 volts. If a $(v/2, v/x)$ select system is used, with the optimum value of x ($\simeq 8$) chosen, ir vidual cells with $3v/8$ across them will, in general, give an out of less than that given in the Table for 120 volts, and will sh a standard deviation which is relatively greater. But there $2(n-1)$ of these cells (i.e. 62 in a 32 × 32 matrix). Thus average total signal will be greater by a factor of 62, wher the standard deviation will be increased by a factor of only $\sqrt{}$ Similarly, the signal from the $(n-1)^2$, i.e. 961, cells with across them would give an average output 961 times, an standard deviation 31 times, that of a single cell. Usin 32 × 32 matrix and an excitation voltage of 240 volts it found that V_u had an average value of 0·27 volt and σ = 0· This means that 95% of the pulses would result in a value V_u of between 0·11 and 0·43 volt, while the probability of being greater than 0·54 volt is less than 0·001.

(These probabilities have been calculated on the assumpt that the voltages of individual pulses have a normal distribut about the mean value. In fact, there appeared to be a sli asymmetry in the distribution when a voltage pulse of amplitude was used.)

It can therefore be concluded that the variation in the li emission from pulse to pulse is unlikely to be a limiting fac in the operation of a 32 × 32 matrix, though it must be allo for in the design.

(6.3) Direct Measurement of V_w/V_u using Masks

To simplify the problem of driving the $2(n-1)$ unselec conductors with fast-rising pulses, a 16 × 16 matrix was u for the experiments. (The capacitance of the 32 × 32 pa was about 0·006 μF, and it was considered that the edge of voltage pulses should rise and fall in less than 0·3 micros

Fig. 13.—Masks used with electroluminescent matrices.

Masks of the three types shown in Fig. 13 were made from t photographic film.

Mask 1 transmitted light from one spot only. By select of the rows and columns of the matrix, light from a cell w V, $(v/2 - v/x)$ or $(2v/x)$ across it could be observed.

Mask 2 allowed light from all the cells except the selec one to fall on the photomultiplier.

Mask 3 allowed only the light from the $(n-1)^2$ cells w voltage of $2v/x$ across them to be observed.

Mask 1 was used to measure V_w, and mask 2 to measure V_u. The total area of the matrix was 2·4 cm square, and a pho multiplier with a cathode 4·4 cm in diameter was used. matrix could not be positioned in close proximity to the cath because of the thickness of the glass plate (see Section 3 It was therefore put at a distance of 5·5 cm. It was found t $b \simeq 142$, and the pulse amplitude used was 136 volts. The fore $b/\sqrt{v} = 12$. V_w/V_u was measured for a number of val of x, and the experimental results have been plotted in Fig. The theoretical curve has been drawn for comparison. It

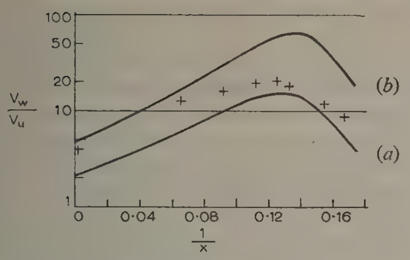

Fig. 14.—Discrimination ratio for 16 × 16 matrix.
(a) $b/\sqrt{v} = 10$.
(b) $b/\sqrt{v} = 12$.
++ Experimental points.

seen that the maximum occurred at a slightly higher value c, and was very much lower than had been predicted. The ilts nevertheless follow a curve which is similar in shape to theoretical one.

he most likely explanation of this lower discrimination is : it is caused by the fraction of the light from the selected which is totally reflected at the glass surface. This light, reaching the lower surface of the glass, will not encounter a s-to-air boundary, and at least part of it will be transmitted he underlying phosphor layer, where it will be scattered and orbed. But some of the scattering will be in a forward ction, and this will result in the escape of light which inated behind an opaque square through holes in the mask. n addition, there is a possibility that a small amount of ismission sideways away from an excited cell can occur, so : some light escapes into the surrounding phosphor, where it / be scattered forward, and may 'leak' out through holes in mask. Such an effect is likely to be confined to within a ll distance from the excited cell.

hat there is an unwanted escape of light was demonstrated the use of mask 3, with $x \to \infty$, i.e. with all conductors hed except the selected pair. There should have been no t reaching the photocathode from the matrix, but an output al of about $0{\cdot}04V_w$ was observed. It was confirmed that was due to light emitted from the matrix by removing the age pulse from the selected conductors; this resulted in the put dropping to zero.

(7) CONCLUSIONS

n the paper the most exacting situation has been assumed. these circumstances a discrimination ratio of 22 : 1 was ained with a 16 × 16 matrix. It can be inferred from the ves in Fig. 9 that a ratio of 9 or 10 : 1 should be possible a 32 × 32 matrix. Even when allowance is made for the dom variation of the output signal, these figures suggest that hould be possible to use a 32 × 32 matrix as a permanent e. With the values of $b/\sqrt{v}$ attainable in practice it is oretically possible to use a 64 × 64 matrix, but with the erimental matrices at present in use the discrimination ratio ıld be too low. Improvements will depend more on reducing scattered light than on improving the ratio $b/\sqrt{v}$. Better ılts may be obtained by making the panels on very thin sheets lass, or by using opaque conductors on the glass and transparent conductors on top of the phosphor. A very thin resin layer may be used to protect the top (transparent) conductors. With a thin transparent layer, much more of the internally reflected light would be returned to the phosphor of the emitting cell or to the phosphor immediately below the opaque border which surrounds each cell.

It has been shown that there is no theoretical limit to the smallness of the cells, but if they are made very small a larger voltage pulse must be used, and this leads to a reduction in the theoretically possible discrimination ratio. Cells which are 1 mm^2 in area have been used, but it is obvious that these do not represent the lower limit in size, and reduction to one tenth of this area could be tolerated.

The fundamental limitation to the speed of the store is the decay time of the phosphor after excitation, or the 'afterglow'. It seems reasonable to assume that considerable improvements can be obtained in this direction.

(8) ACKNOWLEDGMENTS

The authors would like to thank Prof. F. C. Williams and Dr. T. Kilburn for their help and advice in discussing the work reported in the paper.

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DISCUSSION ON THE ABOVE FOUR PAPERS BEFORE THE MEASUREMENT AND CONTROL SECTION, 1ST MARCH, 1960

r. J. E. Flood: The form of store shown in Fig. 1 of the er by Prof. Kilburn and Dr. Grimsdale was also indepentlly invented two years ago by my former colleague, the late R. W. S. Kinsey.

In the store proposed by Mr. Kinsey the coils corresponding to the binary digits of a word were packed, one behind another, on a common axis. The values of the digits were determined by assembling small ferrite cylinders and non-magnetic spacers in a

transparent tube. The tube was then inserted through the centre of the coils in order to provide the required coupling paths between windings. Assemblies of coils were placed side by side to form a row of word stores, and several such rows were mounted one above another in a rack. In this way a 3-dimensional store was achieved, instead of the planar arrangement described by the authors.

A small model of the store was built and successfully demonstrated. However, we did not proceed further with the development, and the reason for this may be of interest. We were investigating semi-permanent stores for use as translators in automatic telephone exchanges. In this application a very short access time is not required; a variety of stores are therefore available, including the magnetic drum and the Dimond ring translator.*

The Dimond ring translator and its successors store information by means of a variable pattern of wires threaded through a fixed array of magnetic cores. The present store uses the inverse arrangement, namely a variable pattern of cores threaded through a fixed array of wires. Because we did not need high operating speeds we failed to see any advantage in the new arrangement and did not develop it further. The authors are to be congratulated on realizing the advantages to be gained from this inversion. As the wires have fixed positions, they can be laid out carefully and kept as short as possible, thus enabling high operating speeds to be achieved. At the very high speeds used by the authors, it would appear necessary to consider the wires as transmission lines. If this is so, information on the parameters of these lines would be useful.

The mechanical design of the changeable version of the authors' store is novel and the idea of moving the ferrite cores by means of air jets is most ingenious. It would be interesting to know how reliable this arrangement has proved in practice.

Mr. C. P. Gerrard: What does the 200 000-bit permanent store cost, and how does it compare with core assemblies? If it is not essential to vary the store, a version using small ferrite rings of linear magnetic material would seem to offer the following advantages:

(*a*) No keeper rods would be required. The store could be made more compact, thus reducing the propagation time and giving a faster read-out.

(*b*) The drive currents would be reduced because of the higher-efficiency magnetic circuit formed by the small ring. This would reduce the time spent in allowing the current to reach an effective value, and thus further reduce the read-out time.

(*c*) One could make use of existing production techniques for threading square-loop cores.

There remains, of course, the question of initial errors in the threaded assembly, which is one of the disadvantages of the system, but a few errors can be dealt with. In the system described, any bit to be read out is uniquely specified by the voltages at two points. One point is in the drive selector and the other is in the read selector. If a bit is in error, these two points can give advance warning that an incorrect bit has been accessed. A purely logical correction can then be effected using approximately three semiconductor elements per incorrect bit. Therefore, a few errors can be tolerated with no great increase in cost, but, of course, no drastic changes can be made.

It would seem, therefoie, that, although it is not so flexible, a system using very small linear ferrite rings would cycle at a higher rate and perhaps be reasonable in price.

Mr. T. H. Walker: I have used small ferrite toroidal cores not for storage, but for coding and code translating.

I agree with the authors about the advantages of using the two-core-per-bit system, but would query the speed. Using fastest currently available materials and employing partial f switching in a word-selection system, the 2 microsec cycle ti seems a little slow. Since we were dealing with much smal matrices, this may be significant, and I wonder whether authors could give some idea of the relationship between total cycle time and the speed of switching of a single eleme

Mention is made of increased temperature sensitivity wh using the two-core-per-bit system. Do the authors agree t the tolerances with respect to core characteristics must be bet for this system, because one is dealing with second-order rat than first-order properties in this case?

The permanent or semi-permanent type of magnetic st obviously represents basically a very cheap method, but I won what proportion of added expense is involved in the mechani for making it possible to change the information, which it wo seem must be of fairly high precision.

Mr. W. A. E. Loughhead: I am interested in the code transla as applied to the telephone field. Dr. Flood mentioned Dimond-ring type of translator, in which wires are fed throu cores. One of the problems encountered was interference fr external magnetic fields, etc. I notice that in the read-out st there are rather long loops of wire. Have the authors had a trouble with pick-up and interference problems on account these loops?

Mr. D. Eldridge: The ability to change the contents of fixed store described in the paper by Prof. Kilburn and I Grimsdale raises the question whether such a facility is re required. There is inevitably a price to be paid for this flexibi if only in that a more economical technique using resistors capacitors would be otherwise practicable. It may well be t the provision of this feature is only justified in either a univers or special-purpose machine or where the store contents can changed rapidly and automatically so as to provide an auxilia input channel for data.

When compared with a random-access core-matrix store fixed store can offer both a speed and cost advantage. Howev the latter is offset somewhat by the selection scheme, which direct rather than coincident, and, with the technique describ by the need to insert a large number of keeper rods.

The 52-sense amplifiers used in conjunction with 832 transis switches have not been described by the authors. Some det of these would be welcome together with an assurance that difficulty is encountered in switching 30 millimicrosec 2½ r signals.

The stated cycle time of 0·2 microsec appears to be obtaina only at the expense of limiting the essentially random-acc nature of the store, since about 1 microsec is required to swi the sense amplifiers. As used by the authors this is acceptab but its use for a control system as described in Reference 2 the paper requires a completely random-access store. T restriction therefore either limits the field of application or speed of this system.

Mr. A. Proudman: With regard to the fast-carry propagati circuit, are there any loading problems on the chain of transist forming the carry path itself?

The authors divide the carry path into groups of six stag with emitter-followers between groups. Transistors associa with the least-significant digits of each group will therefore ha to drive six outputs in the case where a carry is propagati I notice that both the stages shown in the Figure are identic and that no obvious allowance is made for the tee-ing of current down the chain. Is any problem associated w this?

Mr. E. H. Cooke-Yarborough: In any kind of magnetic-c store it is necessary to drive a current down any one of ma

* Dimond, T. L.: 'The No. 5 Crossbar AMA Translator', *Bell Laboratories Record*, 1951, 29, p. 62.

…es. That is true whether the store is of the coincident current … word-selection type.

…here is one way of doing this which I have not heard …ussed, and I wonder whether Prof. Kilburn has considered … It is the selection switch described by Constantine.* If a …ent is to be driven down one of N wires, each wire is coupled …one of N transformers. N pairs of transistors are also …pled to these transformers, every transistor being coupled to …y transformer, but with a different combination of phases …ach case. To select a particular wire, one transistor of each … is energized and their effects are so phased as to cancel …pletely in all the transformers except the one associated with … selected wire, where the effects add. Thus the N transistors … effectively in parallel in driving the selected wire, and quite large currents should be obtainable. I would be interested to learn whether this was tried as a way of driving a square-loop store.

The velocity-of-light carry reminds us of the problems which will arise when computers become so fast that the physical size starts being limited by the velocity of light. Is it essential to locate all the high-speed parts of the computer within, say, one digit period of one another, or it is practical to consider a logical design in which the whole computer is much larger, but is so laid out that there is never a long distance between components which are required to interact? One might think that this would be very difficult to arrange, but I have heard that it would place only minor restrictions on the logical design.

* CONSTANTINE, G.: *IBM Journal of Research and Development*, 1958, 2, p. 205.

THE AUTHORS' REPLIES TO THE ABOVE DISCUSSION

…**rof. T. Kilburn** and **Dr. R. L. Grimsdale** (*in reply*): It is …cult to give useful estimates of the cost of commercial …ions of the permanent store, but it is suggested in the paper … the cost per bit would be between one-tenth and one-…tieth of that for the conventional square-loop magnetic-core …e. The substitution of rings of linear ferrite material for the …s used in the permanent store would seem undesirable as it …ld increase the cost and make changes in information …emely difficult. No trouble is experienced with pick-up in … mesh due to stray magnetic fields.

…he transistor switches used for 'read' selection operate satis-…orily. They have negligible impedance to the 'read' signals …n closed, and when open have an impedance which is high …pared with the input impedance of the read amplifier.

…he store has been arranged so that there are 16 blocks, each …256 words. The access time to a word in a block is … millimicrosec and a block will remain selected until a new … is required. The time to change from one block to another …ess than 1 microsec. As the information is permanent, … library routines, the information can be arranged with-…he store so that block changes are not required within …ines.

…**r. D. B. G. Edwards, Mr. M. J. Lanigan,** and **Prof. T. …urn** (*in reply*): The memory cores used have a switching … of 1·2 microsec when operated under coincident current …ditions. In the basic bias system using word selection and …flux switching the switching time is reduced to 0·42 microsec, …ng a minimum possible cycle time of 0·84 microsec. This …e will be increased in practice mainly owing to the time required to establish the operating currents. If transistors are to be used to drive a 1 024-word store, this latter time could well be 0·6–0·7 microsec. Thus the cycle time is of the order of 1·5 microsec, the figure of 2 microsec being essentially a target in the initial specification of the store requirements.

Using partial-flux-switching techniques either the cycle time or the power requirements of the system can be reduced.

The tolerance of a system to variation in core characteristics depends on the exact details of the system. There are some two-core-per-bit systems which are just as sensitive to variations as a one-core-per-bit arrangement; in particular, the coded system which has been referred to in the main text. It is therefore impossible to agree with the general statement made by Mr. Walker. With regard to temperature sensitivity, we have observed that a two-core-per-bit full-flux-switched system was less sensitive in this respect than a similar system with partial-flux switching.

Prof. T. Kilburn, Dr. D. B. G. Edwards, and **Mr. D. Aspinall** (*in reply*): Mr. Proudman is quite correct in his observation that no allowance is made for teeing in of current down the 'carry' path of the adder. It was felt that in an engineering design such individual adjustment could not be allowed. We have solved the problem by defining the currents in the base and emitter leads fairly accurately, so that in each stage the difference current is a maximum of 0·25 mA. In a group of six stages the total difference current is 1·5 mA and the voltage change is typically 0·1 volt. The purpose of the emitter-follower after a group of six stages is to give a fresh start to the following group with regard to both the difference current and change in voltage level. This technique has proved very satisfactory in practice.

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AN EXPERIMENTAL TRANSISTOR-CONTROLLED COMPONENT SELECTION AND TESTING MACHINE

By T. C. CARDWELL, B.Sc.(Eng.), Graduate, J. R. W. SMITH, M.Sc., A.Inst.P., and G. H. KING, B.Sc., Associate Member.

(*The paper was first received* 7*th September*, 1959, *and in revised form* 1*st January*, 1960. *It was published in April*, 1960, *and was read before the* MEASUREMENT AND CONTROL SECTION 26*th April*, 1960.)

SUMMARY

The paper shows the need for a selection and testing machine in the light of recent developments in the automatic assembly of components on to printed-wiring boards. The advantages of a programmed machine are given and the flexibility and simplicity obtained are illustrated.

The general operation of the machine is outlined. Examples of the method of coding and the technique of programming on to punched-paper tape are included. The build-up of the electronic control from a number of relatively simple basic circuits is shown.

Finally, the range of components handled, details of the test bridges, and the overall speed and reliability are given.

(1) INTRODUCTION

The advent of the transistor and methods of mounting components on boards in a 2-dimensional array offer the prospect of assembling these components automatically, thus saving considerable human effort. The boards or cards used are normally made of phenolic resins and have electrical conductors, usually of thin sheet copper, bonded to one side. Holes are punched or drilled in the card and components are inserted with the ends of the wires bent over on to the copper foil. Components, held in magazines, are assembled sequentially and it will be appreciated that, should an incorrect component be inserted owing to an error in value, tolerance or sequence, the location and rectification of the fault could be both time-consuming and expensive, particularly as the components must be soldered into position before functional tests can be performed. It is always possible for a component magazine to be incorrectly filled and although most components are correct when made, there is no guarantee that they are still so when required for use after what may be a considerable storage period. The need for a check of the components immediately prior to automatic assembly is thus clearly indicated.

A very wide range of electronic equipment may be produced by these methods, involving a considerable variety of components and a large number of different designs. A testing machine should be capable of checking any of the components used and should be so made that it can be changed over easily and rapidly from one design to another. This dictates extreme flexibility and therefore a programmed type of machine has been evolved. The testing sequence forms part of the programme; thus, not only is the machine made simpler by the omission of sequence control, but also the type and order of test can readily be changed. To fulfil these requirements, contributions are needed from the logical designer as well as from the mechanical, electrical and measurement engineers. Furthermore, the conflicting requirements from these diverse fields must be reconciled.

Although the required functions are complex, a deliberate policy of using the simplest and most reliable techniques h been adopted. As will be seen, this has resulted in the use some well-known devices. This approach enables the electror control for such an industrial machine to have the requir degree of reliability without a large-scale proving programme every individual part of the circuits.

The authors are with British Telecommunications Research, Ltd.

(2) GENERAL DESCRIPTION

The machine has been designed to handle axial-lead co ponents only, since these comprise the great majority of comp nents in use to-day.

An outline drawing of the experimental machine is shown Fig. 1. Each component type is mounted in a removal

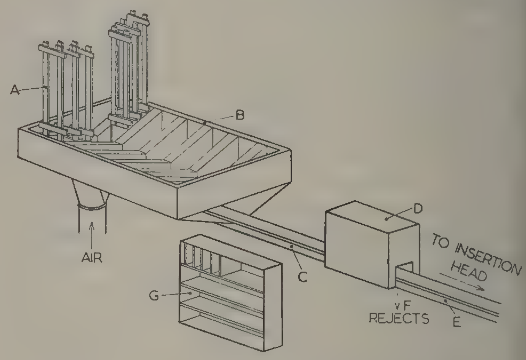

Fig. 1.—Outline of machine showing main parts.

magazine, A, the required number of these being mounted in removable framework, B, comprising the component stor When an appropriate solenoid on the framework is energize a component is released and drops on to the component tran port bar, C. Slits in this bar are fed with compressed air a the components are lifted by the air jets and carried to the te station, D. Here the component is held, test clips make electric contact with it and tests are made. Correct components a permitted to continue along the transport bar at E whilst incorre ones are dropped through the trapdoor, F.

The operation of the test station is shown diagrammatically Fig. 2. The whole sequence is controlled by electronic circui mounted at G in Fig. 1. For ease of construction and mainte nance the circuits are broken down into small units, each mounte on a printed-wiring board which slides into the shelves as show Five-track punched-paper tape, well known in telegraph equip ment, is used for the programme which is read into the machin by means of a step-by-step tape reader presenting one charact at a time. The programme consists of information regardi magazine numbers, component types (resistor, capacitor, diod etc.), component values and tolerances, punched on to the tape

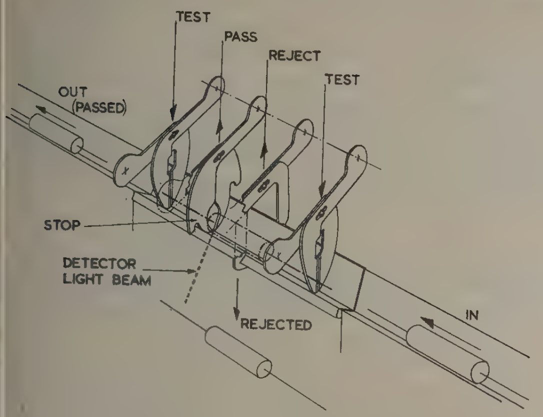

Fig. 2.—The test station.

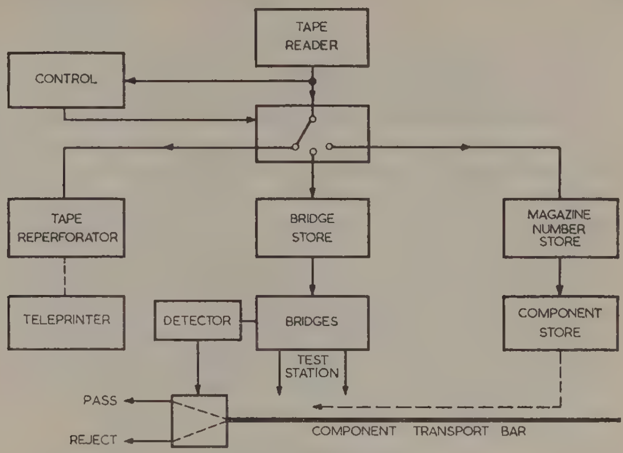

Fig. 3.—Block schematic.

ıps of characters or blocks. Each block of information s with a prefix character which identifies it, and is followed n end-of-block character.

ne whole control system is arranged to reset whenever an of-block character is received. The prefix of the next block en used to route the following information to the appropriate of the machine, and as soon as one block has been dealt the next block is read in. Blocks giving magazine numbers, ponent values, etc., which are required for some time, are ed. The timing of all operations is controlled by pulses ved from a clock consisting of several monostable circuits nected in a ring. Some of the information, after storage, is to set up the required bridge standards by means of reed ys. The pass or reject traps are operated by solenoids. to-transistors are used to detect the movement of com- ents; one is used in the test station to initiate the test cycle he arrival of the component; others detect the passage of pted and rejected components.

(2.1) Outline of Operation

block schematic of the machine is shown in Fig. 3. A cal sequence of events when the programme tape is inserted the reader and the machine started is as follows. The tape er consists of a number of end-of-block characters which are in and used to reset all control toggles. The first block nformation on the control tape is a prefix, followed by the ling or board number and any other special instructions. ng on the prefix, the control routes the information through switch shown in Fig. 3 to a tape reperforator. A copy of heading on the programme tape is thus available for subse- nt use by the progress control or other sections of the factory. next block of information is a second prefix and the number he magazine in which the component first required is stored. control routes this information to the magazine number e and uses it to pulse the appropriate magazine electromagnet, allowing the first component to fall on to the transport em. A third block comprises a prefix and the bridge rmation. This is routed to a common bridge store and is used to set up the range, value and tolerance of the appro- te bridge. In the meantime, the component is being carried he transport system to the test station. On its arrival, it is cally detected. Test clips grip the component's wires and nect them to the test bridge. The electrical test is carried and the component is passed or rejected as appropriate. If rejected the control will call up another component accord-ing to the information still stored in the magazine-number store and the test cycle will be repeated. When a component is finally passed, the control cancels all stored information and steps the tape on to the next position. The whole sequence is repeated for the second and subsequent components until the board is completed.

(2.2) Codes

The codes used are shown in Table 1 in which a '1' is used to represent a hole, and a '0' to represent no hole. Of the

Table 1

Odd Parity Codes

Track number 1 2 3 4 5	Combination number	Meaning as a prefix	Meaning as a tolerance
1 0 0 0 0	0	Spare	Spare
0 0 0 0 1	1	Tape reperforator	±5%
0 0 0 1 0	2	Magazine	±10%
1 0 0 1 1	3	Spare	Spare
0 0 1 0 0	4	Resistance	±20%
1 0 1 0 1	5	Capacitance	
1 0 1 1 0	6	Stop	
0 0 1 1 1	7		
0 1 0 0 0	8		
1 1 0 0 1	9		
1 1 0 1 0	10	Spare	Spare
0 1 0 1 1	11		
1 1 1 0 0	12		
0 1 1 0 1	13		
0 1 1 1 0	14		
1 1 1 1 1	15		
0 0 0 0 0	End-of-block character		

Track 1 is the parity track. The weights associated with tracks 2, 3, 4 and 5 are 8, 4, 2 and 1 respectively. The feed holes lie between tracks 2 and 3.

2^5 combinations possible in each row of 5-track tape, half have an odd number of 1's, and half have an even number of 1's. Those combinations with an odd number of 1's are known as odd parity codes, and are numbered 0 to 15. Any number from 0 to 15 may be written in binary form as $8A + 4B + 2C + D$, where A, B, C and D are 0's or 1's. A is always punched in track two, B in track three, C in four and D in five. Track one

contains a '0' or '1', so that the number of 1's in each row is odd, and it is known as a parity track. This is an example of a weighted code, so called since each track (except track one) has a value or weight associated with it.

With the exception of the code 00000, which always means 'end of block' each of the tape characters may have various meanings according to its position within any block of information. For example, in a block of information consisting of the code numbers 43152, the prefix 4 implies that 3152 concerns a resistance. The 3 is used to indicate that the value lies between 10^3 and 10^4 ohms, and the 15 are the first two significant figures of the value, while the 2 gives the tolerance as $\pm 10\%$. Coding for capacitances is similar and the block 55204 is the code for $2 \cdot 0 \times 10^5$ pF $\pm$ 20%. Magazine blocks are even easier to code, magazine 25 being represented by 225. After the prefix 1, any set of characters will be copied by the reperforator until the code 00000 is encountered. Finally, the block consisting simply of the prefix 6 is used to stop the machine.

(2.3) Programming

As the preceding examples have shown, it is fairly simple to write down and read codes for machine control. By writing first a magazine block, and then the type, value and tolerance of the component contained in the magazine, then another magazine block and so on, the machine is caused to select and test components in any desired sequence. 'Comments' can be interpolated for reperforation, and the tape ends with a stop code. From the written programme, a tape is punched utilizing a suitable keyboard perforator. By comparing the tape with another produced from a second independently written programme, errors may be detected and by tape-editing processes a correct tape can be prepared. No special difficulty has been experienced in preparing tapes for experimental and testing purposes, but more sophisticated programming methods could be introduced if the scale of operation justified them.

(3) THE ELECTRONIC CONTROL

Some of the parts from which the machine is built up will now be described. Positive logic is used throughout, i.e. a '1' is represented electrically as a voltage positive with respect to earth and a '0' as a zero (or negligible) voltage.

(3.1) The Transistor Toggle Circuit

For storage purposes the bistable Eccles–Jordan or flip-flop circuit shown in outline in Fig. 4, and sometimes called a

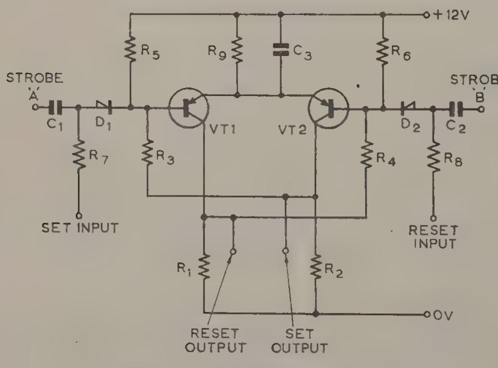

Fig. 4.—Transistor toggle circuit.

R1, R2 2·2 kΩ
R3, R4, R5, R6 10 kΩ
R7, R8 56 kΩ
R9 1 kΩ
C1, C2 0·005 µF
C3 0·01 µF
D1, D2 OA95
VT1, VT2 OC71

'toggle' circuit, is widely used in the control equipment. T circuit can be switched into one of two states by the applicat of control potentials to the 'set' or 'reset' inputs, the change-o being initiated only at the instant of the timing or strobe pul Thus, if many such circuits are required to change they will start to do so during the same strobe pulse, all circuits hav reached their new states before the next strobe pulse occu This technique prevents faulty operation and ensures that machine keeps in step.

(3.2) The Monostable Circuit

Another frequently-used circuit is a monostable version of toggle circuit. The monostable circuit (Fig. 5) requires a cont

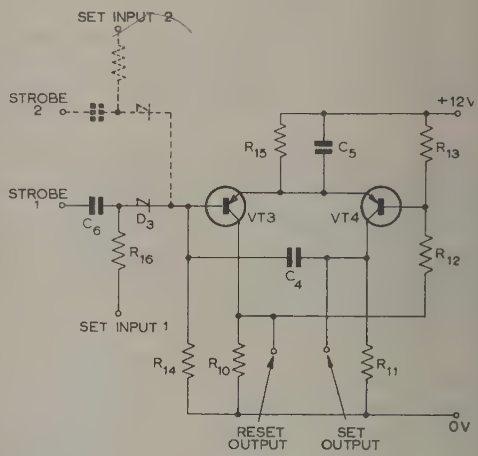

Fig. 5.—Monostable or delay circuit.

R10, R11 2·2 kΩ
R12, R13 10 kΩ
R14, R16 56 kΩ
R15 1 kΩ
C4 as required
C5 0·01 µF
C6 0·005 µF
D3 OA95
VT3, VT4 OC71

potential at the 'set' input to make it respond to a strobe pu in a similar way to the toggle circuit, but since the 'set' state unstable the circuit returns to the original or 'reset' state after time determined by the time-constant of the circuit. A seco independent input circuit is shown in broken lines, allowi operation from either of two sources as required.

(3.3) Diode Gates

Another basic unit is the diode gate. The 'and', or coinciden gate consists of a number of diodes connected as shown Fig. 6(*a*). These gates are used for routing 5-wire informati to the various parts of the equipment, a typical 5-wire gate bei shown in Fig. 6(*b*).

The 'or' gate is also extensively used and is simply a dev for combining signals without allowing any interference betwe sources, Fig. 6(*c*).

(3.4) Sealed Reed Relays

A sealed reed relay has been used in places where transist are unsuitable. This relay consists of two nickel-iron reeds w gold-plated contacts, sealed in a glass tube. It is closed whe current flows in a coil surrounding the tube, this energizi current being switched by a transistor. These relays are id for switching bridge circuits, especially if a screened-coax switch is required. They can also be used to switch volta too large for present-day transistors.

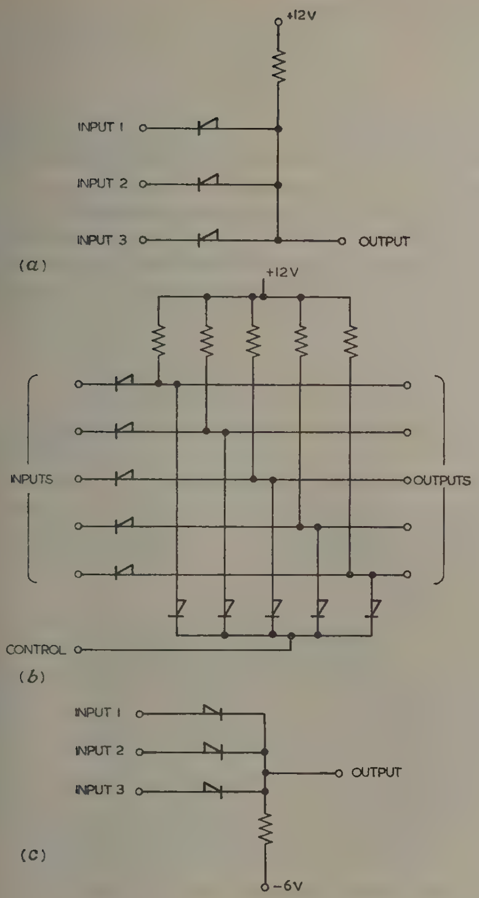

Fig. 6.—Gate circuits.
(a) 'And' gate.
(b) 5-wire gate.
(c) 'Or' gate.

(3.5) Stores

3oth the bridge information and the number of the magazine ıtaining the required components have to be stored. The :dge store consists of toggle circuits arranged in groups of ; for storing the 5-unit characters as they are fed in from t: paper tape. The toggles are set up according to the pattern ·holes in the tape. The tape information is routed to each ɔup of toggles in turn by means of diode gates, as previously ·scribed. After the test has been performed and the component ·cepted as satisfactory, the store is cleared by applying a control ·tential to the 'reset' inputs of all toggles so that the next :obe pulse will change them to the 'reset' state.

(3.6) The Control Timing Cycle

In order to provide strobe pulses and tape-reader drive pulses c·rectly spaced in time, a clock consisting of a number of mono-: ble circuits is used. Fig. 7 shows these circuits arranged in a r g so that a continuous train of pulses can be obtained from ay point. On removing the potential from control 1, the ring i broken and the operation stops. It can be restarted by a·plying a potential to control 2 and injecting a 'start' pulse, c·ntinuous cycling being maintained by the re-application of the ·tential to control 1. Two tape-reader drive pulses, A and B, a shown in Fig. 7, separated in time by 10 millisec. Pulse A

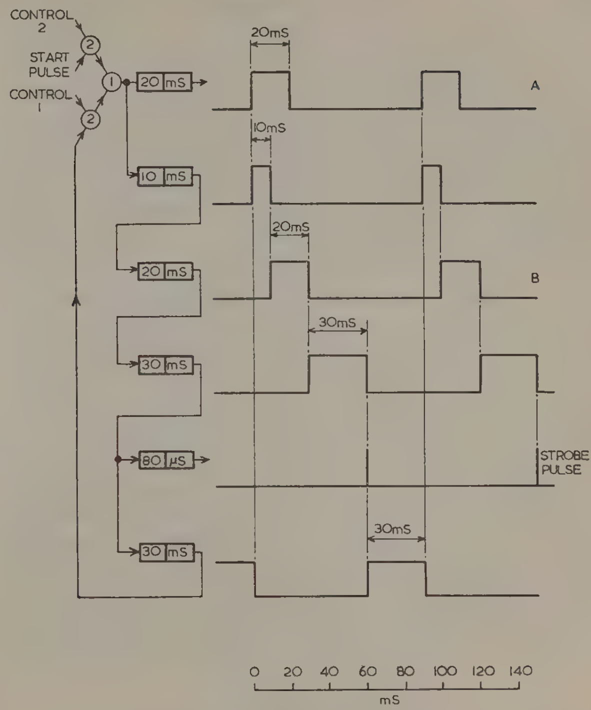

Fig. 7.—Control timing cycle.

causes the tape-reader drive fingers to engage with the tape and pulse B causes forward movement. The termination of A causes disengagement from the tape and termination of B causes the return of the driver. Having moved the tape on by one character, the 'peckers', or contacts, are given time to settle down before the strobe pulse causes the new character to be read.

The tape-reader mechanism is operated by electromagnets which are energized by transistor-controlled current pulses, as shown in Fig. 8. As considerable power is necessary to drive the tape-reader magnets, several stages of current gain have been used.

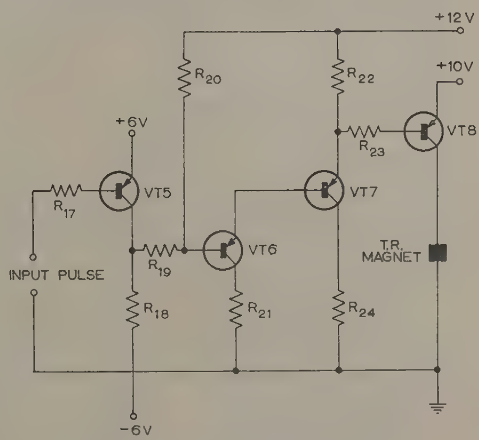

Fig. 8.—Tape-reader drive circuit.

R17, R19 10 kΩ	R23 10 Ω
R18 1 kΩ	R24 75 Ω
R20 3·9 kΩ	VT5, VT6 OC71
R21 2·2 kΩ	VT7 OC72
R22 22 kΩ	VT8 OC16

(3.7) The Translator

The tape codes shown in Table 1 are used directly by the bridges but require translation to a series of single-wire outputs for use in bridge selection, magazine selection, etc. Translation is performed by a diode matrix, as shown in Fig. 9. According to the state of the tape-reader contacts, various combinations of high and low potentials (1's and 0's) are applied to lines 1 to 5. By means of five transistor inverters (only one is shown) the inverse combinations are applied to lines 6 to 10. The outputs A to H are effectively the outputs of 'and' gates and

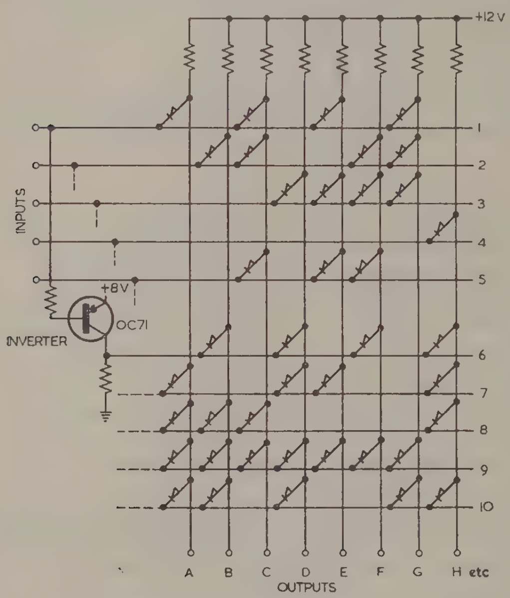

Fig. 9.—Translator matrix.

each gate will give a positive output when, and only when, corresponding input combination is applied.

(3.8) Interconnections

The output from the tape reader is connected permanently the input of the translator. The tape information is theref available at all times in 5-wire code and single-wire form. previously explained, a series of 5-wire gates are used to ro information to the bridge store. Each gate is opened and sh by its particular control toggle. The bridge prefix code alrea mentioned gives an output from the translator which is used cause the several gate-control toggles to operate in sequence a so to read each character into the appropriate part of the sto Similarly, another prefix causes the magazine information (af translation) to be stored.

Information regarding the progress of components through test machine is given by a number of photo-transistor opti detectors, this information being used in conjunction with di logic circuits and monostable toggles to control the timing the individual events in the testing cycle.

Finally, two toggles, which are set according to the result of test (i.e. 'pass' or 'reject'), are used to drive, by means of am fiers similar to that of Fig. 8, the solenoids controlling the 'pa gate and 'reject' trap.

(4) COMPONENTS AND TESTS

The present transport system will not allow very small very large components to be handled, but resistors of $\frac{1}{4}$ w and larger and paper capacitors up to $0{\cdot}25\,\mu$F are possil Improvements in the transport system and re-designed test c would increase this range.

A simplified diagram of a capacitance bridge is shown Fig. 10. This differs from normal bridges by having all switch operations performed by reed relays and having standa arranged in binary decades. Thus the value of capacitance be measured is set up on the standards in binary decimal fo

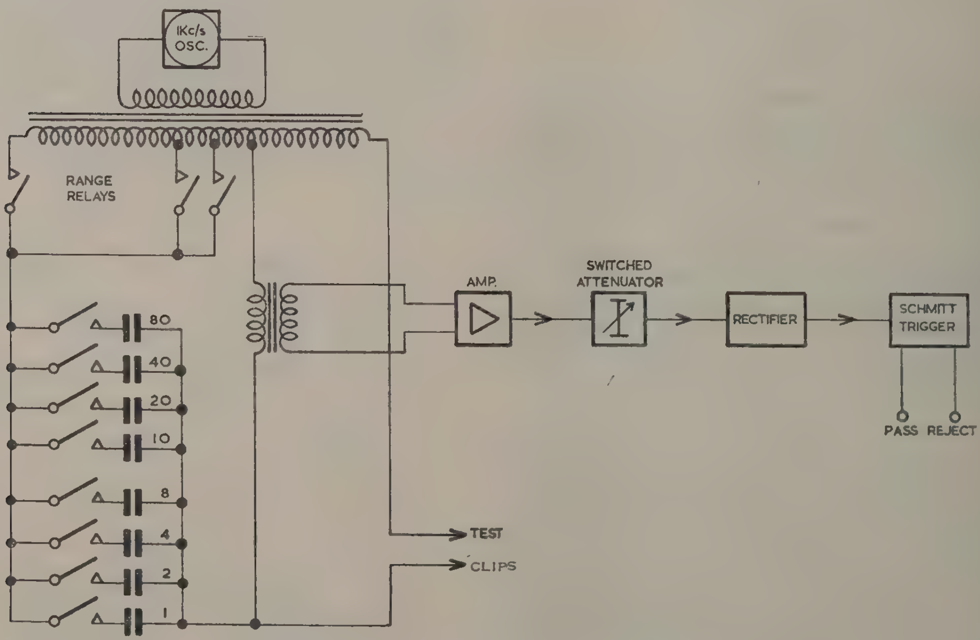

Fig. 10.—Simplified circuit of capacitance bridge.

rectly from the third and fourth characters of the block of formation supplied by the tape, the appropriate range switch eing set according to the second character of the block of formation. An example of this coding is given in Section 2.2. he bridge output is amplified and passed through a switched tenuator, the value of which is set by the tolerance information fth character). With a given deviation of component value om nominal the rectified bridge-output signal is arranged to t the state of a Schmitt trigger circuit.

The resistance bridge is similar to the capacitance bridge, but e standards consist of a series chain of resistors with switches ross them, these being opened according to the resistance value pplied by the tape.

Values catered for by these bridges are 100 ohms to 100 kilohms d 0·001 to 10 μF.

(5) SPEED AND RELIABILITY

The testing machine is part of a component assembly plant pable of handling many thousands of components a day. It therefore essential that both speed and reliability shall be fully lequate. In fact, the testing time is only a few milliseconds d consists mainly of the time required for the test clips to me into contact with the component. The main restriction is the time of transport of the components and this limits the ze of component store that can be used, owing to the increased stance to be covered with larger stores. In practice, the time r component transport and testing has been arranged to be ss than the cycle time of the component assembly machine. ith this restriction a store for 50 component types has been hieved.

Occasionally, owing to dirt on the leads or other causes, bad ntact is made by the test clips on to the component. When is happens the component is automatically rejected and a fresh one is called up. Reliability is of a high order at the cost of the occasional rejection of a good component.

(6) CONCLUSIONS

The need for flexibility has been fully substantiated and the use of a paper-tape programme to achieve this has proved an economical solution to the problems involved. Even so, a considerable quantity of information must be stored and logically processed for the operation of this machine. In the interests of reliability and freedom from maintenance it was decided that the minimum number of moving parts would be used, and results with the experimental machine fully confirm this decision. The use of strobed-toggle techniques to read and store information and diode gates to direct operating signals to the various parts of the machine has also been found to be both reliable and flexible. It has been possible to get the whole of the control and logic into a volume less than 1 ft^3. Relays of conventional design were considered for the logic circuits but were rejected on grounds of bulk and the number of contacts involved. In those places where metallic contacts could not be avoided, such as the test bridges, sealed reed relays have been used with complete success. These relays are reliable, have a long life and can be driven from transistor circuits. The combination of transistors, diodes and reed switches has been found very suitable for this type of programmed testing equipment, being robust, reliable, easily maintained and capable of rapid modification when required.

(7) ACKNOWLEDGMENTS

We wish to thank our colleagues for their help and assistance, and the Director of Research, British Telecommunications Research, Ltd., for permission to publish the paper.

DISCUSSION BEFORE THE MEASUREMENT AND CONTROL SECTION, 26TH APRIL, 1960

Mr. J. A. Sargrove: I was concerned in the development of earlier automatic component-testing equipment (Acte) which s designed with relay-operated logic since at that time trantors were less reliable than they are now. Its purpose was evaluate 2-terminal components for type approval after very curate life testing.*

In this machine there is a large jig for the 1 000 components d 100 motorized 10-way switches which constitute one large 00-way 2-pole switch. This enables the loading and/or asuring circuits to be connected to each component in turn, value being measured and recorded.

The other 999 components not being measured at the time n continue to be loaded with alternating current so that their bient temperature conditions do not materially alter while e individual component is being tested.

There are built-in bridges and an unusual 1 000-chart recorder, h chart displaying the sorted information on the life history each component.

There is the usual logical system for operation and interpretan of the results, all circuits being constructed with thermionic ves and relays.

I believe that Acte was the first punched-tape-commandntrolled testing machine built. The purpose of this method command is to pre-programme automatically, say, 100 days life testing. Commands for all sorts of climatic cycles can punched and the climate-producing equipment (heaters, rigerators, humidifiers, etc.) are made to change or maintain desired environmental condition within the testing cubicle without the need for human intervention. This liberates scientific personnel for more important work.

The film 'First Automatic Factory' (made in 1947) shows a high-speed (five per second) hopper and inspection and pass-reject sorting system. A noteworthy feature is the use of a statistical method of automatically orienting the asymmetrical component automatically in the inspection station. This was well before its time, and even now it is seen but rarely. It is worthy of further development as being capable of helping many automation schemes.

Mr. F. G. Swift: At first consideration, a printed-wiring board is only an alternative form of wiring. Its true virtue, however, lies in promoting a new form of assembly. This opens the way to a radical change in the whole conception of design, manufacture, inspection, production control, presentation of engineering information, ordering, stocktaking, and accountancy. All these can now be integrated by common-based data and the gains stemming from such an integrated manufacture are very considerable compared with those due to printed wiring considered only as an alternative form of wiring.

Consider, for instance, the production of engineering information. Two-dimensional printed-wiring boards designed for automatic component insertion demand the adoption of modular principles of layout. If the co-ordinate points of component terminals and fixings together with board dimensions can be stated modularly, dimensionally accurate drafting is not required and the demands on a drawing office in terms of skill and effort can be reduced or, indeed, removed entirely by the adoption of photographic techniques, or by a simple tabulation of co-ordinate points, clerical fashion or as holes in a tape.

* SARGROVE, J. A.: 'Automation in Component Testing', Report of International posium on Electronic Components, Malvern, 1957, p. 127.

Tape control of manufacture and inspection provides as a by-product data in a form ready for processing. Exploiting this to the full, production control and costing is presented with a new and powerful tool.

Will the authors amplify their reasons for adopting sequential reading of information within a block, and stores and logic circuits, when the results could be achieved by reading a static block of information directly from the tape via relays?

In the bridge measurements, a null balance is not obtained but the level of the out-of-balance signal is related to the deviation from nominal of the component value. Will the authors describe the steps taken to stabilize this relationship?

Mr. A. H. Coppard: Are the authors going to make any provision for testing transistors, since the machine seems rather limited owing to the lack of that facility?

Is any trouble experienced due to contact resistance on the clips, which could cause resistances having a low negative tolerance to be passed? Finally, can a bad power factor in a condenser affect the measurement of its capacitance owing to the alteration of the unbalance signal by the in-phase component, and thus permit condensers which are out of tolerance to be passed?

Mr. W. Fordham Cooper: Some attention must be given to the diagnosis of faults, possibly after a substantial period of use. In the past, if a component gave trouble one was able to refer to a still legible record in the drawing office. The idea of having to refer to a punched tape which is in several parts is rather daunting. An apparatus can function correctly for quite a long time and then fail in an emergency because of an undetected error in the connections, or of an unsuitable or inaccurate component. There may also be a minor but fundamental mistake in detail of the design. With automatic selection and assembly of components everything depends on the punched tape. It is possible to guard against fingering errors in typing by duplicating the work and checking one tape against another, but there is still the possibility that the engineer may have made a slight error in the programme, and an error on a punche tape would not be nearly so obvious as an error in a drawing In a recent conversation with young engineers responsible fo development of logical circuits, they could very easily tell m what would go right, but I was left in considerable doubt whethe they had thought sufficiently deeply about what could go wrong This is, in fact, a very much more difficult problem in circui design.

The arrangement of the apparatus we have seen may not b the best for more complicated assemblies, and the authors wi find, when they come to designing machines for such purpose that it would be better to test the components separately at th magazine before they reach the assembly line, rather than havin one test point carrying out a number of different tests accordin to signalled instructions. I think that the best arrangemen however, would be for tests to be carried out as the magazine are being charged, with only an approximate check test befor assembly, not to ensure that the individual component is of th right magnitude, but merely to check that it has come from th right magazine. This follows what is, I think, an importan principle in sequential design, namely that all checks an measurements should be made at the earliest, rather than th latest, convenient stage. This greatly simplifies logical desig and layout and is of importance in designing servo mechanism and control systems.

Mr. J. K. Webb: Fears about the effects of human program ming errors have been expressed, and it should be realized tha these arise in all data-processing systems. The present syster appears to be a relatively simple one and such errors should no give much trouble if the usual precautions are taken. It ha been found that, with two trained girl operators employed t punch and verify tape, the residual error rate is only about 1 i 30 000. Special programming techniques can be evolved fo dealing with the majority of these residual errors, and what sti remain can usually be coped with by a little common sense o the part of the supervisors.

THE AUTHORS' REPLY TO THE ABOVE DISCUSSION

Messrs. T. C. Cardwell, J. R. W. Smith and **G. H. King** (*in reply*): We find ourselves in substantial agreement with Mr. Swift. To gain the maximum advantage from the introduction of tape-controlled machines requires considerable re-thinking and changes to all parts of the production and design organizations. Our choice of a tape reader was decided partly by the lack of a proven block reader, and partly by the limitations inherent in the fixed block length associated with such readers. The total information concerning one component cannot be contained within the available space of known block readers, and the need to allow for addition to or reduction of the information required dictates the use of a flexible system which the block reader cannot provide.

The method of using a bridge in an out-of-balance manner is not satisfactory without careful stabilizing of the oscillator output amplitude. We no longer use this method but prefer a digitally balanced bridge giving a near-null final setting. In this case a phase-sensitive detector is used and signal amplitudes are not so important. This obviates the problem mentioned by Mr. Coppard regarding wrong test decisions being taken on capacitors with bad power factors. Also in answer to Mr. Coppard, the assembly machine used in conjunction with compo nent selection and testing equipment does not insert transistor and consequently the provision of transistor test equipment di not arise. The adequate testing of transistors is, howeve another field in which automatic methods can be of considerabl value and we are currently engaged on such work.

Mr. Cooper has made some very interesting points, especiall regarding the assembly of circuits containing slight inaccuraci of design. In our view this is adequate justification for leavin the accurate measurement of component values until the late convenient stage, provided that a record of the actual value each component in a specific board is made. This is a powerf tool for use during the development and early production stag of a circuit.

Some of Mr. Cooper's comments on tape errors are answere by Mr. Webb. However, we may add that the codes have bee carefully chosen so that, when the tape is printed out by mea of a teleprinter, the record obtained is easy to read. Experien with tape tends to convince us that it is in fact easier to chec tapes, using the tape comparison methods now available, tha to compare a printed record or drawing directly.

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The Institution of Electrical Engineers
Paper No. 3294 E
Nov. 1960

THE RECORDING OF HIGH-SPEED SINGLE-SHOT PHENOMENA

By F. E. WHITEWAY, B.Sc., Associate Member.

(*The paper was first received* 11*th November*, 1959, *and in revised form* 25*th April*, 1960.)

SUMMARY

Techniques at present in use for the recording of single-shot phenomena by cathode-ray oscillographs are reviewed. Cathode-ray tubes are discussed, with particular attention to the conflicting requirements of high writing speed, wide bandwidth and high sensitivity. These requirements can be met to a large extent by the use of very small spot size and travelling-wave techniques. An account is given also of associated circuits including high-speed time-bases, amplifiers and calibration equipment.

(1) INTRODUCTION

The problems associated with high-speed recording by a cathode-ray oscillograph are particularly acute for single-shot phenomena, since the oscillograph must possess a very high writing speed. Measures taken to increase the writing speed usually result in a loss of deflection sensitivity, making the amplifier problem particularly difficult when high sensitivity and wide bandwidth are required. Sampling techniques cannot be used to solve this problem. The methods at present in use for recording single-shot high-speed transients are discussed in this paper, particular attention being given to the conflicting problems of high writing speed, wide bandwidth and high sensitivity. A greater emphasis has been placed on sealed oscillograph tubes than on continuously evacuated tubes, since the former are far more convenient to use in recording systems and are now being produced with very high writing speeds. Complete recording systems are not considered, since these depend on the particular application. However, the design and performance of many of the individual units of the recording systems are considered, including amplifiers, time-bases, time calibrators and pulse generators. The application of direct-viewing storage oscillographs has not been included, being at the present time restricted to writing speeds of 10^8 trace widths per second or less.

(2) CATHODE-RAY TUBES

(2.1) Factors Governing Spot Size

The cathode-ray tube (c.r.t.) is the principal component of an oscillograph, and sets a limit to the overall performance that is possible. For repetitive phenomena, measurements can often be made directly on the tube face using a graticule, but for single-shot phenomena the trace must be recorded photographically. In this case the physical size of the recording area is unimportant over a wide range, provided that adequate resolution can be achieved. The resolution depends on the maximum number of traces which can be contained in a horizontal or vertical direction within the recording area, such that each trace can just be distinguished from the next. Normally at least 100 trace widths are required in each direction. Trace width is a function of both spot size and trace speed, whilst spot size is dependent on beam current. The size of the record can be adjusted by altering the magnification of the lens, or by enlarging after development. A limit is reached when the spot size becomes comparable with the finite grain sizes of the phosphor and photographic emulsion, and with the resolution of the lens. This limit is normally approached at spot diameters of about $25\,\mu$.

The design of cathode-ray tubes with very small spot size has been considered in detail by Haine and Jervis.[1] A higher deflection sensitivity can be achieved, as defined in terms of volts per trace width. Also an improvement can be obtained in the maximum writing speed of the tube, defined in terms of the maximum number of trace widths which can be recorded per second for a given optical system and film. Examples of oscillographs with very small spot sizes are Lee's micro-oscillograph,[2] Haine and Jervis's demountable oscillograph[3] and the sealed K.R.3 travelling-wave c.r.t.[4,5] These have spot sizes quoted as 10μ, 40μ and 38μ respectively. It is significant that all employ electromagnetic focusing, and that in two cases the focusing lens is situated well forward in the vicinity of, or even beyond, the deflection system.

(2.2) Factors governing Maximum Writing Speed

The maximum writing speed of a display system is of fundamental importance for high-speed recording. It depends mainly on the energy of the beam, the current density of the spot, the type of phosphor, the efficiency of the optical system (if any) and the nature of the emulsion. The last two factors are considered in Section 3, but it should be pointed out that they are interdependent to some extent. Factors affecting current density are considered by Haine and Jervis.[1] High current density requires high beam current and small spot size, and it is current density which accounts for the difference in writing speed between available high-sensitivity tubes and those of moderate sensitivity using the same phosphor and high-voltage supply (e.h.t.) (cf. tubes A and B, Table 1). Whilst there is an appreciable difference in sensitivity between these two types of tube in terms of volts per centimetre, the difference is less marked in terms of volts per trace width. Furthermore, the size of the recording area in terms of trace widths is greater in tube B.

The maximum writing speed of a c.r.t. increases with the energy of the beam, i.e. with the overall e.h.t., although the relationship is not linear because the phosphor efficiency also increases with e.h.t.[6] In order to avoid low deflection sensitivities resulting from the high overall potentials, most high-speed c.r.t.'s employ post-deflection accelerators (p.d.a.'s), so that the potential at the point of deflection relative to the cathode is much lower than the overall potential. The accelerators are usually in the form of bands round the bulb of the c.r.t. for low p.d.a. ratios (defined as the ratio e.h.t./gun potential). For high p.d.a. ratios it is necessary to use a spiral or graded p.d.a. to avoid raster distortion. At least one type of cathode-ray tube has been developed with a p.d.a. ratio as high as 15 : 1. This uses a screen to prevent penetration of the field of the p.d.a. into the deflection system, which would result in severe raster distortion. In general, an overall e.h.t. of at least 10 kV is required.

This is an 'integrating' paper. Members are invited to submit papers in this category, giving the full perspective of the developments leading to the present practice in a particular part of one of the branches of electrical science.

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

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Finally, the maximum writing speed of the c.r.t. is dependent on the type of phosphor used. The efficiency of the phosphor is affected by the energy of the beam, and two phosphors of equal sensitivity at a given potential may be quite different in sensitivity at other potentials. The relative efficiency of two phosphors is also dependent on the speed of the transient when the duration of the excitation pulse is much less than the decay-constant of the phosphor. This is considered by Feinberg,[7] who shows that the efficiency falls as a result of the significance of non-radiative dissipation of energy at high spot velocities. Another important factor in the choice of phosphor is in matching its spectral response to that of the emulsion. This means in general that a blue or green phosphor is required. The blue P11 silver-activated zinc-sulphide screen with a short decay-constant is widely employed, being in general superior to the green P1 manganese-activated zinc-silicate phosphor. The efficiency of a phosphor is dependent on its preparation and also on the subsequent c.r.t. manufacturing history. Most phosphors are backed by a thin film of aluminium to increase the forward efficiency, to prevent damage to the phosphor by positive ions and to provide a low-impedance return path for the electrons.

(2.3) Bandwidth of Deflection System

Another important factor in a high-speed c.r.t. is the bandwidth of the Y-deflection system. When simple deflector plates are used the finite transit time of the beam through the plates results in a loss of sensitivity at high frequencies, due to the change in phase of the signal. This is considered in detail by Lewis and Wells,[8] who show that the sensitivity is reduced in the ratio

$$\frac{\sin \frac{1}{2}\omega t_r}{\frac{1}{2}\omega t_r}$$

where t_r = Transit time of beam through the plates.
$\omega = 2\pi \times$ signal frequency (f).

The -3 dB point occurs at $ft_r = 0{\cdot}422$, so that for $t_r = 1$ nanosec (4 cm long plates with 5 kV on gun), $f = 422$ Mc/s.

The series inductance of leads and the internal capacitance also limit the high-frequency response by forming a tuned circuit, the resonant frequency of which is often called the 'plate ringing frequency'. This is also considered by Lewis and Wells, although the circuit cannot always be treated as consisting of a single inductance and a single capacitance in series. For wide bandwidth it is essential that the transit time, and the inductance and capacitance of the plates and connecting leads, should be ke as small as possible. Lee[2] carried this to an extreme by usi a very high voltage (50 kV), short plates (0·2 in long), ar coaxial connections. He estimated the bandwidth of the syste to be approaching 10 Gc/s. In spite of the high voltage ar short plates the sensitivity was 10 volts per trace width, which comparable with more conventional tubes (type B, Table 1).

By combining micro-oscillograph techniques with travellin wave deflection systems, very wide bandwidths may be achiev coupled with high sensitivity. The KR3 travelling-wave c.r (Fig. 1), made by Edgerton, Germeshausen and Grier of Bosto

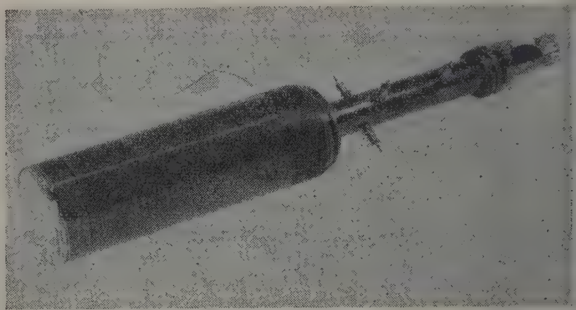

Fig. 1.—The KR3 travelling-wave cathode-ray tube.

Mass., has a bandwidth of 2 Gc/s with a sensitivity of 30 n for a deflection of one trace width (0·0015 in). A descripti of the early development is given by Germeshausen *et al.*[4] T travelling-wave deflection assembly (Fig. 2) contains two defl tion lines, each consisting of a flat helical conductor wound insulators supported by a central earthed metallic strip. T beam passes between the two helices, and each alternate ha turn of each helix acts as a separate deflector plate of very sh transit time, namely 0·1 nanosec. The velocity of the sig down each deflector line is matched to the velocity of the bea and the effective transit time of the deflector line corresponds that of the single turn. Thus long lines can be used, and si a very small spot diameter has been achieved, the result sensitivity is high. The helices are matched in characteris impedance to external coaxial cables, and the signal may be through them to other equipment if required. Wider ba widths have been achieved with other, though less sensiti

Table 1

Details	Cathode-ray tube type					
	A	B	VCRX 434	Demountable	TW11 (Ref. 9)	KR3
E.H.T. on gun, kV	2	5	7·5	40	10	4
Overall e.h.t., kV	10	10	30	40	35	22
Method of focusing	Electrostatic	Electrostatic	Electrostatic	Electromagnetic	Electromagnetic	Electromagnet
Trace width, mm	0·5	0·33	0·33	0·04	0·10	0·038
Phosphor	P11	P1 or P11	P11	None	P11	P11
Max. writing speed, trace-width/sec	2×10^9	2×10^{10}	2×10^{11}	5×10^{12}	$3{\cdot}5 \times 10^{11}$	10^{11}
Y-sensitivity, volts/trace-width	0·35	1·5	1·8	Not given	0·46	0·03
X-sensitivity, volts/trace width	1·5	2·0	2·0	2·0	0·80	0·20
Y-plate ringing frequency, Mc/s	400	440	700	—	—	—
Y-plate transit time, nanosec	1·9	0·9	0·6	—	0·086/turn	0·1/turn
Bandwidth, t.w. system, Mc/s	—	—	—	10 000*	approx. 3 000	2 000
Impedance, t.w. system, ohms	—	—	—	—	125	120
Max. Y-deflection, trace-widths	80	300	200	Not given	340	270
Max. X-deflection, trace widths	200	300	200	Not given	1 000	400

* Estimated bandwidth at 30 kV when using parallel-wire deflection system.

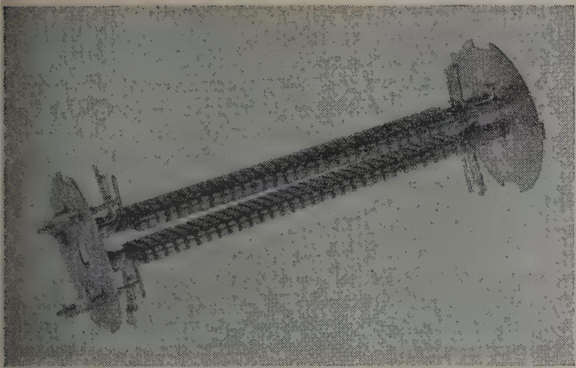

Fig. 2.—Deflection assembly of KR3 tube.

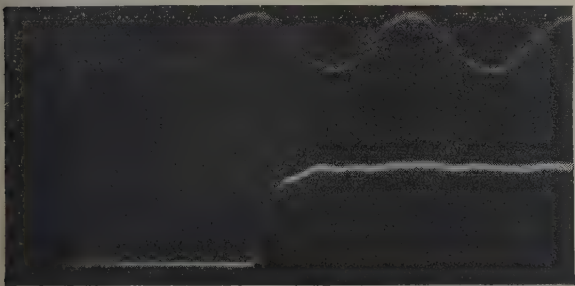

Fig. 3.—Pulse displayed on KR3 tube.

travelling-wave tubes.[9] Fig. 3 shows the pulse from a relay pulse generator displayed on a KR3 travelling-wave c.r.t. using coupling cables with a delay of 30 nanosec. The calibration waveform is 300 Mc/s, from which it can be seen that the pulse recorded has a fast edge of about 0·5 nanosec. The slow rise following the fast edge is believed to be a function of the coupling cables.

The signal to be displayed does not always require an oscillograph of high sensitivity. Advantage can then be taken of using a very high accelerating potential for the c.r.t., thus increasing the writing speed. Furthermore the length of the deflector plates may be reduced appreciably in order to reduce transit time. In the limit a parallel-wire deflection system at right angles to the beam may be used.[10] This has the advantage of wide-band transmission characteristics as well as very low transit time. As the pulse potential is increased still further some form of high-voltage divider becomes necessary. These are often of a resistance-capacitance form, or a pure capacitance form, and have been considered in detail by Howard.[11] Dividers of the capacitance type may be incorporated inside the c.r.t., and a c.r.t. suitable for the display of 100 kV pulses has been described.[12]

(2.4) Spiral-Trace Oscillograph Tubes

When high timing accuracy is required, the length of a single trace may not be sufficient in terms of trace widths, since measurements can only be made to an accuracy of about one-quarter of a trace width. A raster type of display may be used to increase the total length of the trace. Alternatively a spiral trace may be employed. Special c.r.t.'s have been developed for this purpose, such as the G.E.C. type 1358B, in which the X- and Y-plates are both used as time-base plates and radial deflection is achieved by applying the signal between a cone mounted on the faceplate and a conductive coating on the side of the bulb. Displays of this type have been designed at the Atomic Weapons Research Establishment (A.W.R.E.) with up to 20 turns in the recording area, giving a maximum trace length of about 500 cm (15 000 trace widths).

(2.5) Currently Available Cathode-Ray Tubes

Table 1 gives a comparison between some currently available c.r.t.'s. It will be noticed that sensitivity is given in terms of volts/trace-width. This is sometimes called the 'sensibility' or 'figure of merit' of the c.r.t. Examples of tubes having a performance similar to type A are: Tektronix type T54P11 and G.E.C. type 5BHP11. Examples of type B are: 20th Century Electronics type S5AB/123 (P11 phosphor) and Electronic Tubes type 5BVP1 (P1 phosphor). At the time of writing several manufacturers are developing c.r.t.'s of comparable writing speed and sensitivity to the VCRX434 and with similar e.h.t. requirements. The demountable tube of Haines and Jervis[3] is given for comparison. The writing speeds given should be taken only as a guide, since they do not refer to the same recording system in each case. The figure given for types A, B and the VCRX434 apply for a Wray F/1 lens and R60 or HPS film.

(3) PHOTOGRAPHIC RECORDING[6,13,14]

(3.1) Choice of Lens

The camera, together with its associated lens and recording film, is an important component of the recording system. The design of the lens is of particular importance, since it can affect the resolution and maximum writing speed of the system, and can introduce pattern distortion. It must have a short focal length for compactness, and also a wide aperture for high-speed recording, i.e. a low f-number. The effective aperture of the lens is given by $(f/d)(m+1)$, where f is the focal length, d the effective diameter and m the magnification of the lens. Thus m should be as small as possible, although little will be gained by reducing it below $\frac{1}{4}$. Furthermore, large reduction ratios may result in the resolution being impaired by the grain size of the emulsion. The Wray $f/1$ copying lens is used extensively for high-speed recording. This lens is designed for a fixed magnification of $\frac{1}{4}$, and thus has an effective aperture of $f/1{\cdot}25$; it has a focal length of 2 in, and is bloomed for blue-green light; it is suitable for a recording area of 4 in diameter, using 35 mm film. Tests at A.W.R.E. have shown that the lens is capable of resolving lines spaced 0·002 in apart. Germeshausen *et al.*[4] recommend the use of a 1 : 1 optical system for c.r.t.'s with very small trace widths, such as the KR3. Two Dalmeyer $f/1{\cdot}5$ lenses may be used in tandem to give an effective aperture of $f/1{\cdot}5$ at a magnification of unity. The resolution and transmission are satisfactory only over a very restricted area (about $\frac{1}{2}$ in diameter), but this is sufficient for the KR3 tube, which has a small recording area. Very good results have been obtained over a similar area using two Canon $f/1{\cdot}2$ lenses in tandem at a magnification of unity. For general purposes, when maximum speed is not required, the Wray $f/2$ lens has been found to be very versatile. It is satisfactory over a wide range of magnification, and the depth of focus is much less critical than with the Wray $f/1$, since the ratio of the effective lens diameters is 1 : 2·9. The depth of focus is obviously a function also of trace width, since a greater degree of defocusing can be permitted for a wide trace than for a narrow one. Taking the maximum permissible diameter of the spot produced on a film by a point source on the phosphor to be one-quarter of a trace width, the depth of focus is equal to $ub/2d$, where

u = Distance of c.r.t. phosphor to front node of lens.
b = Trace width.
d = Effective diameter of lens.

With a Wray $f/1$ lens the depth of focus is thus $(9 \cdot 2/4)0 \cdot 33$ mm, i.e. $0 \cdot 76$ mm for a trace width of $0 \cdot 33$ mm, and only $0 \cdot 09$ mm for a trace width of $0 \cdot 038$ mm.

(3.2) Choice of Film

There are a number of factors governing the choice of suitable photographic emulsion. Hopkinson[6] has pointed out that two emulsions which have identical sensitivity under steady conditions may have quite different sensitivities for transients of short duration. Also, as stated previously, the spectral response of the phosphor and emulsion should be matched as far as possible. This is more likely to govern the choice of phosphor, since for all photographic emulsions the spectral sensitivity is greatest at the blue and violet end of the spectrum. Kodak R60 and Ilford HPS film developed for maximum contrast with Kodak D19B or Ilford ID19 developer are both satisfactory for high-speed recording. Recent tests using Agfa Record film in conjunction with May and Baker's Pentalex developer have been very promising. When maximum speed is not required, a lower-speed film with finer grain is preferable.

(3.3) Cameras

The choice of camera will depend on many factors. When single transients are to be recorded in a laboratory, plate cameras may be preferred, since there is no possibility of the size of the record being altered during processing or storage. For most purposes, single-shot or continuous-feed cameras using 35 mm film are satisfactory. Continuous-feed cameras may be obtained covering a wide range of speeds. At higher speeds it is important that the film does not move more than one-quarter of a trace width during the decay time of the phosphor.

(4) LINEAR AMPLIFIERS AND ATTENUATORS

(4.1) Cascade Amplifiers[15]

The design of amplifiers for high-speed transients has received much attention during recent years. The complexity of the amplifier required to feed the deflection system of an oscillograph will depend on three main factors, namely bandwidth (or rise time), output voltage and gain. For moderate speeds, cascade amplifiers can be designed using valves possessing a high 'figure of merit'. This is a measure of the quality of the valve for high-frequency operation and may be interpreted as the frequency at which the gain is unity when the input capacitance C_g and output capacitance C_a form the load. The figure of merit is therefore given by $g_m/2\pi(C_a + C_g)$, where g_m is the mutual conductance of the valve. There have been substantial advances in the production of suitable valves in recent years, using fine grid wires and close cathode-grid spacing.[16] A number of manufacturers are producing valves with figures of merit of at least 200 Mc/s, and there are several types of secondary-emission valve which have figures of merit of this order. The combination of the modern close-spaced valve as a primary stage with at least one stage of secondary emission would result in even higher values. The gain-bandwidth product of a stage is constant for a given form of coupling, and for a simple resistive load this product is equal to $g_m/2\pi C_T$, where C_T is the total shunt stray capacitance. Let this value be represented by B. Compensation circuits may be used to improve the gain-bandwidth product, and by using valves with equal input and output capacitances, constant-K low-pass filters may be used to couple stages together.[17] The maximum product that can be achieved theoretically is $4B$, although in practice a value of $2 \cdot 8B$ represents a realistic maximum. Unfortunately these circuits are very critical in the distribution of the stray capacitance. It is therefore more common to use some simp form of high-frequency compensation, such as the insertion an inductance in series with the load resistor. The gai bandwidth product for this simple circuit is about $1 \cdot 5B$ f critical damping. It should be noted that the use of feedbac such as in cathode compensating circuits, does not improve t gain-bandwidth product. From the above it will be seen th low-level cascade amplifiers with bandwidths of 100 Mc/s a feasible. However, many c.r.t.'s require about 200 volts to gi a deflection of 100 trace widths. This means that if the to output capacitance is 20 pF, then for a rise time of 10 nanos the output valve must pass a current of at least $0 \cdot 4$ amp order to charge the capacitance alone. If the valve is requir to amplify information of either polarity the quiescent curre and power dissipation will be too high for existing wide-ba valves, even if a push-pull system is adopted. For informati of one polarity only, the output valve may be driven positi from a low quiescent current, using feedback to linearize t amplitude response. Secondary-emission valves may be used give a symmetrical output, and may also be used in the precedi stages, coupling from dynode to grid to avoid phase inversio All valves can then be operated at high electrode potentials a low steady currents, so that they can deliver large current a voltage pulses. An amplifier of this type has been designed wi an output of 300 volts, a gain of 300 and a rise time of und 20 nanosec for a mean h.t. current drain of only 80 mA.

For very-high-speed single-shot transients, the use of pre-pul techniques may be considered.[18, 19] This method requires t application of a pre-pulse of about 50 microsec prior to t signal in order to switch the amplifying valves into conditions high current and high voltage. The increase of current a potential results in a substantial increase in figure of merit the type of valve used (type EFP60), owing to an increase mutual conductance from 23 mA/volt under normal conditio to 130 mA/volt under pulse conditions. As a result it is possib to design very linear cascade amplifiers with wide bandwidt and high output voltages. For example, one amplifier has bandwidth of 140 Mc/s and an output of 300 volts peak-to-pe for a sine wave input of 1 volt peak-to-peak. It uses only fo amplifying valves. Wider bandwidths can be achieved for low output swings.

(4.2) Distributed Amplifiers

It is standard practice to use distributed techniques wh designing amplifiers of bandwidth in excess of 100 Mc/s, a distributed amplifiers are often used for output stages abo 30 Mc/s. The design of such amplifiers has been considered many authors.[8, 20, 21] When valves are connected directly parallel there is no increase in the figure of merit of the combin tion, since both the mutual conductance and stray capacitan are increased in proportion to the number of valves. Howeve by connecting the valves in such a way that the input and outp capacitances form the shunt elements of two separate distribute lines, the bandwidth to a first order is dependent only on t line parameters and not on the number of valves. Thus t required gain may be achieved by connecting sufficient valves parallel, even if the required bandwidth exceeds the figure merit for each valve. Many types of line can be used, but t one most commonly adopted is that shown in Fig. 4(a). It an m-derived transmission line with a series inductance L p section, shunt capacitance C per section (stray capacitance and mutual inductance M between adjacent sections. The u of a suitable degree of mutual coupling between sections improv the phase response of the line, and a virtually constant del can be achieved for frequencies up to $0 \cdot 6f_c$, where f_c is the cut-o frequency of the line.

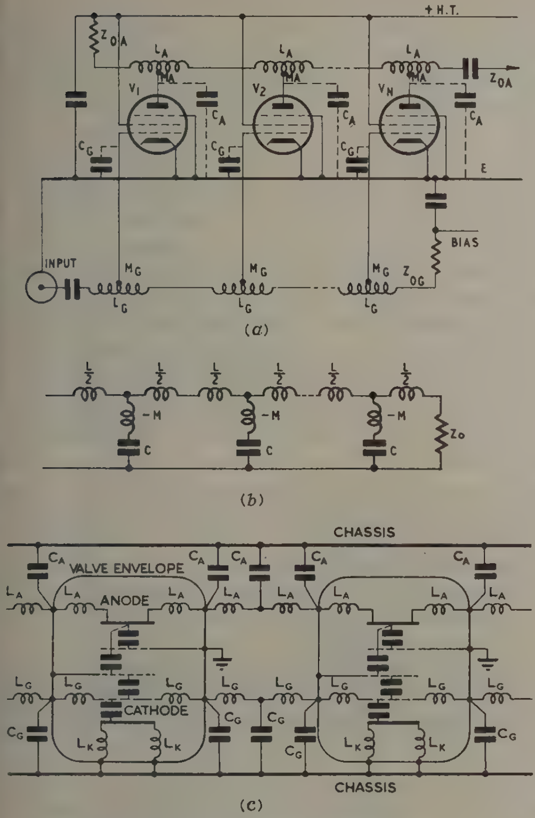

Fig. 4.—Distributed amplifier.

(a) General circuit.
(b) Equivalent circuit of m-derived line.
(c) Use of dummy sections.

The optimum values given by Lewis and Wells[8] are

$$m = 1{\cdot}27$$

Coupling coefficient $k = -0{\cdot}23$

Characteristic impedance

$$Z_0 = 1{\cdot}11\sqrt{(L/C)}$$

Delay per section $= 1{\cdot}11\sqrt{(LC)}$

Cut-off frequency $f_c = \dfrac{0{\cdot}36}{\sqrt{(LC)}}$

Fig. 4(b) shows the equivalent circuit of the line, with a negative inductance term in series with the shunt capacitance. Any lead inductance in series with the shunt capacitance will result in a reduction in the effective mutual inductance between sections and hence in a poorer phase response. It is sometimes possible to increase the coefficient of coupling to compensate for this, but as the bandwidth is increased the value of the line inductance becomes smaller and it may not be possible to obtain sufficient mutual inductance to offset the lead inductance alone. When these conditions are approached, the use of a separate input and output lead to each electrode may be advantageous. In this case the lead inductances are in series with the line inductance and not with the shunt capacitance. In the limit the line inductance of each section may consist solely of the inductance of a single internal lead using dummy sections between the valves and a capacitance to earth at each electrode terminal [Fig. 4(c)]. The Westinghouse Company (U.S.A.) are advertising valves of matchbox construction designed specifically for distributed amplifiers, which have input and output leads for each electrode brought out on opposite sides. Two amplifiers each with a gain of 3 have been built in A.W.R.E. using the above techniques. One of these uses experimental pentodes with two grid leads and two cathode leads. The other amplifier uses cascode-connected triodes, which have two grid leads and four cathode leads. In the grid lines the series inductance consists solely of the stray inductance of the input and output grid leads with flat strips of low inductance connecting one valve to the next. The anode lines of these amplifiers have a bandwidth of 500 Mc/s, although the grid lines restrict the overall bandwidth to about 360 Mc/s. The grid line usually presents the greatest design difficulty in a distributed amplifier, because of the attenuation produced at high frequencies by the input conductance of the valves, which is proportional to the square of the frequency over a wide range. Although there are methods for reducing this effect, such as the inclusion of an inductance in series with the screen lead or by use of negative feedback from the cathode, the number of valves and the complexity limit most amplifiers at the present day to an upper frequency limit of about 300 Mc/s. This is shown by a recent survey.[22] At 200 Mc/s, low-signal amplifiers can be designed with comparative ease, but the output stage is likely to present a considerable problem if linear outputs of about 200 volts are required, because of the high dissipation incurred. The use of a sensitive high-speed c.r.t. such as the KR3 is thus a great advantage, since it requires only about 10 volts for full deflection.

(4.3) Attenuators

When the amplitude of the signal to be displayed is too large in comparison with the voltage required for full deflection, some form of attenuator will be required. A useful review of wide-band attenuators is given by Lewis and Wells.[8] Wide-band fixed attenuators for insertion in low-impedance transmission lines are usually constructed from series rod and shunt disc elements mounted in coaxial containers.[23] As may be expected, the bandwidth that can be obtained with attenuators is far in excess of possible amplifier bandwidths. With reasonable care, fixed attenuators can be designed to cover a frequency range up to 1 000 Mc/s. Switched attenuators are usually restricted to the order of a few hundred megacycles per second.

(4.4) Coaxial Systems

In many oscillographs where the time-base is triggered from the signal, it is necessary to insert delay in the signal channel to counter delays in the time-base and beam-brightening circuits. Sometimes there is sufficient delay in the amplifier (if used), but in many cases it is necessary to use delay cable. For oscillographs with rise times exceeding 50 nanosec, delay cables with impedances of the order of 1 000 ohms may be used to give delays of up to 1 microsec with little effect on pulse shape, provided that both ends are properly terminated. For very-high-speed recording the provision of a delay of only 100 nanosec by low-impedance cable can cause quite a serious distortion of the front edge of the pulse. This is illustrated by Fig. 5, which shows the front edge of a very fast pulse after transit through 30 nanosec and 130 nanosec cables. It is thus very important in transient-recording systems with rise times of nanosecond order to keep the length of cable in the recording channel to a minimum. A fact perhaps not generally realized is that the rise time introduced by a cable is proportional to the square of the cable length over a very wide range.[24] The design of wide-band terminations is also important, and for a very-wide-band response it is necessary to make them form part of the coaxial

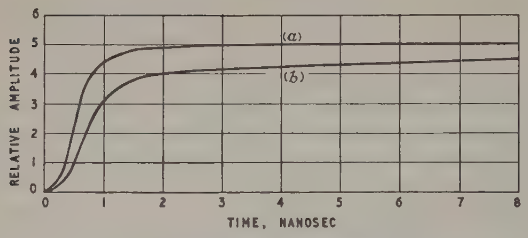

Fig. 5.—Pulse distortion caused by transit through cables.
(*a*) 30 nanosec delay.
(*b*) 130 nanosec delay.

system. Film resistors of the rod type may be used as the inner conductor to form a termination. The outer conductor should then be tapered in the manner indicated by Fig. 6, so that the line impedance at any point matches the remaining resistance to earth. Elementary considerations indicate that the contour of

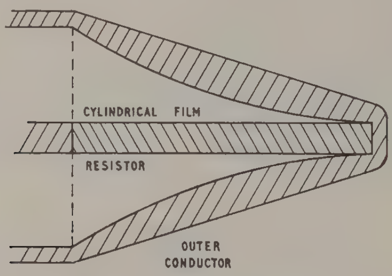

Fig. 6.—Coaxial termination.

the outer conductor should follow an exponential law. However, Harris[25] has shown that the contour required is a tractrix, the difference arising from the fact that the wavefronts are not plane inside the termination. Tractrix terminations can be designed which are satisfactory from direct current to 4 000 Mc/s. However, the design is very difficult to manufacture accurately, and approximations to the tractrix can be used if a more restricted frequency range can be tolerated. Other components in the coaxial system, such as plugs and sockets, also require very careful design when the bandwidth exceeds 500 Mc/s.

(4.5) Wide-Band Transformers

A frequent requirement in a display system is to invert the signal, or to provide a balanced signal from an unbalanced one. Conventional pulse transformers may be used, but it is often very difficult to obtain at the same time both a good low-frequency response and a good high-frequency response. This is because the leakage inductance and stray capacitance form a tuned circuit, and reduction of leakage inductance often entails reduction in primary inductance, or an increase in interwinding capacitance. Toroidal cores using very thin tape of high-permeability material are often used in such transformers to provide a high coupling coefficient. However, pulse inversion may be made in a coaxial system simply by breaking the cable and connecting the inner of each cable to the outer of the other. A balanced output can be achieved by feeding the signal from a cable of impedance Z_0 into two parallel cables each of impedance $2Z_0$, using the inverting technique on one cable only. The above method of pulse inversion is suitable only when the duration of the pulse is shorter than the electrical length of the terminating cable, but if long cables are used the high-frequency response will suffer. Furthermore, the system suffers from bad screening propertie[s]. Considerable refinements can be made to extend the frequen[cy] bandwidth and improve screening. In one system a quarte[r-] wavelength cavity is used which enables a good shield to [be] employed without short-circuiting the signal. The low-frequen[cy] response is governed by the inductance of the outer casi[ng] enclosing the cavity, and may be extended by loading the cavi[ty] with ferrite material;[26] at the same time this broadens t[he] response of the cavity. Another technique[27] uses a twin-wi[re] transmission line wound round a toroidal core of high pe[r-] meability. This device is really a transformer in which t[he] primary and secondary windings are bifilar, and in which t[he] interwinding capacitance forms part of the transmission lin[e.] The high-frequency response is determined by the characteristi[cs] of the line, whilst the low-frequency response is governed by t[he] primary inductance as in a normal transformer. Either type [of] transformer can be designed with a frequency response extendi[ng] from a few megacycles per second up to nearly 1 000 Mc[/s.] Recent developments in A.W.R.E. by H. T. V. Foster (not y[et] published) have extended the frequency to 1 500 Mc/s, with [a] pulse droop of only 10% at 10 microsec.

(5) WIDE-BAND LOGARITHMIC AMPLIFIERS

Some display systems are required to cover a large dynam[ic] range. This may be achieved by using a multiplicity of line[ar] displays staggered in sensitivity. The number of displays ma[y] be reduced if a logarithmic system is adopted in which t[he] amplitude of the signal recorded bears a logarithmic relationsh[ip] to the input signal. The signal amplitude on the face of t[he] c.r.t. can usually be measured to an accuracy of $\pm\frac{1}{4}$ trace widt[h,] i.e. $\pm 0{\cdot}25\%$ for a deflection of 100 trace widths. For a redu[c-] tion in amplitude of one decade the accuracy is $\pm 2{\cdot}5\%$; for [a] reduction of two decades, $\pm 25\%$. However, although [a] logarithmic system covering several decades will give a low[er] accuracy at maximum deflection than a linear one, this accura[cy] remains constant over the whole logarithmic range. It is feasib[le] to design a logarithmic system with a maximum deflection [of] 100 trace widths and covering 4 decades with a constant accura[cy] of $\pm 2{\cdot}4\%$. Four linear displays would be required to cover th[e] range for the same minimum accuracy.

At low speeds there are a number of logarithmic devices whic[h] can be used. An example is the ME1404 electrometer triode, [in] which the anode current has a controlled logarithmic relationsh[ip] to the grid current for at least three decades of positive gr[id] current, i.e. from 3×10^{-12} amp to 3×10^{-9} amp. These a[re] not suitable for wide bandwidths, however, and Fig. 7 illustrat[es] the type of amplifier in use by A.W.R.E.[28] It consists of [a] cascade amplifier of n identical stages, using valves with a hig[h] figure of merit. The polarity of the signal supplied to all stag[es] is the same, and the output is fed through a separate isolati[ng] stage to a common addition line. The operation of the amplifi[er] is as follows.

Let v_i = Input voltage.
v_0 = Output voltage.
A = Gain per stage.

Then for very-low-amplitude signals the output voltage is pr[o-] portional to the input voltage, and is given by

$$v_0 = kA^n v_i\left(1 + \frac{1}{A} + \frac{1}{A^2} + \ldots + \frac{1}{A^{n-1}}\right)$$

where k is a constant.

If the stage gain, A, is at least equal to 3, say, then the ampl[i-] tude of v_0 is mostly governed by the last stage. When the inpu[t]

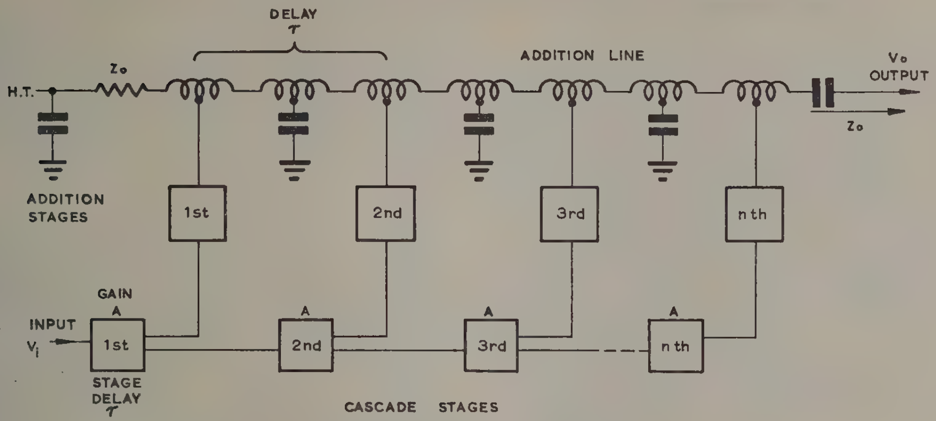

Fig. 7.—Logarithmic amplifier.

signal has risen sufficiently, the last stage saturates, and the output supplied by it to the addition line is constant, equal to V, say. When the input signal has risen by a further factor A the next-to-last stage saturates. The rth stage will saturate when

$$A^r v_i = V$$

so that

$$r = \log_A V - \log_A v_i$$

The output is, to a first order, proportional to the number of stages which have saturated, i.e. $(n - r + 1)$.

Thus

$$\begin{aligned} v_0 &= k_1(n - r + 1) \\ &= k_1(n - \log_A V + \log_A v_i + 1) \\ &= k_2 + k_3 \log v_i \end{aligned}$$

In the above expressions, k_1, k_2 and k_3 are constants.

It will be noticed that a transmission line has been used to couple the addition stages. This must be designed to have the same signal delay per stage as the cascade amplifier. Logarithmic amplifiers of this type with a bandwidth of about 50 Mc/s and covering three decades have been in use by A.W.R.E. for several years. Development is nearing completion of logarithmic amplifiers using a small 200 Mc/s distributed-amplifier strip for each cascade stage. The logarithmic range at these wide bandwidths is restricted by noise, but a range of 3 decades can be obtained. Stability is also a great problem, since there is an inherent feedback path between output and input stages.

(6) HIGH-SPEED TIME-BASES AND TRIGGER CIRCUITS

(6.1) Time-Bases

Time-bases of the bootstrap and Miller type are widely used for high-speed recording for time-base durations in excess of 100 nanosec. Below 100 nanosec the bootstrap time-base often requires a high current in order to obtain a sufficient rate of rise of potential, even to charge the stray capacitance alone. As the speed is increased, non-linearity is introduced owing to the decrease in efficiency of the cathode-follower, and also to the transit time of the feedback loop. In the Miller time-base, one of the main limitations is that the feedback capacitor becomes too small in relation to circuit strays. For very high speeds, circuits relying on feedback to linearize the trace should be avoided. An inductance may be used to maintain the linearity of the trace.[19] The time-base circuit consists simply of a valve passing a steady current and with a tuned LC circuit in series with the anode. The time-base is initiated by cutting off the valve, and the anode potential rises at a rate determined by the original current and circuit capacitance, C. The time-base is, in fact, the first part of a sine wave, and a linearity of 1% can be achieved provided that L is at least equal to $50\tau^2/C$, where τ is the time-base duration. It is necessary to clamp the potential at a suitable level in order to prevent excessive voltages from being produced at the electrodes. The time-base suffers from the disadvantage of requiring high quiescent current, although in some recording systems it is possible to pre-pulse the valve slowly into current, thus reducing the mean current to a low value.

Another type of time-base relies on the constant-current characteristic of a tetrode or pentode in order to maintain the linearity of the sweep.[8, 29] One form of this circuit is illustrated by Fig. 8. It consists of a fast blocking-oscillator driver stage

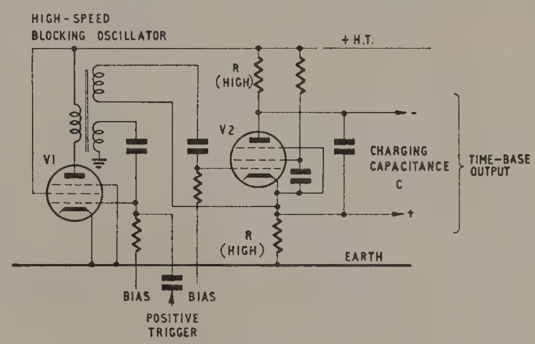

Fig. 8.—High-speed driven time-base.

with the tertiary of the blocking-oscillator transformer connected between grid and cathode of the time-base valve. The charging capacitor is connected between anode and cathode of the time-base valve, and when the valve is driven into a high constant current it delivers a symmetrical time-base output. Time-bases of the type described may be designed for sweep durations of a few nanoseconds. When speeds of this order are required the rise time of the driving pulse often affects the linearity at the beginning of the trace, and it may then become necessary to apply X-shift so that the non-linear part of the time-base does not appear on the screen. The beam current is not switched on during this period. The extra delay required in the signal channel is only in the nanosecond region.

(6.2) Trigger Circuits

The time-bases normally require a preceding trigger stage. At moderate speeds a monostable multivibrator will fulfil this function and also provide the necessary beam-brightening pulse to switch on the beam of the c.r.t. At higher speeds, secondary-emission trigger stages of the type described by N. F. Moody are widely used.[30] These trigger stages employ positive feedback from dynode to control grid or from anode to cathode. Rise times of the order of 5 nanosec can be obtained, and the same valves may also be used to supply the beam-brightening pulse. High-speed blocking-oscillator circuits using transformers with high-permeability toroidal cores can be designed to give similar rise times coupled with high outputs. Transistors are becoming popular for use in high-speed trigger stages as higher-frequency types become available. Transistors operating in the avalanche mode are particularly useful, since they can deliver a pulse with a rise time of the order of 1–2 nanosec with an amplitude of about 5 volts, and of the order of 10 nanosec for an amplitude of 50 volts.[31] Similar results can be obtained with 4-layer diodes.

(7) PULSE GENERATORS AND CALIBRATION EQUIPMENT

(7.1) Pulse Generators

In order to check the step response of a high-speed recording system it is necessary to have a pulse generator with a very fast rise time relative to that of the system being measured and a flat top for the duration of the slowest time-base. It is not always possible to fulfil both of these requirements in the same pulse generator. Pulse generators employing mercury-contact relays to discharge cables[32] are now in general use, and, by mounting the switch to form part of the coaxial system, it is possible to obtain rise times of the order of $\frac{1}{4}$ nanosec. This type of pulse generator is not suitable for long-duration pulses because the length of cable required is impracticable and the back edge of the pulse is deteriorated. A pulse generator which has been designed in A.W.R.E. represents a compromise between the requirements for short rise time and wide pulse width. The mercury-switch contacts are connected in the grid circuit of a high-slope triode, which is driven from cut-off into high current when the switch closes. Pulses with rise times of less than 2 nanosec are generated at the anode and cathode of the triode. The anode-grid capacitance of the triode gives rise to a small transient of opposite polarity in the anode waveform which arrives before the main pulse, as a result of the finite transit time of the valve. Neutralization is necessary to eliminate this effect. The pulse is terminated by a high-speed trigger stage, operated by a variable delay circuit. The resultant pulse width is variable from 0·2 to 200 microsec, with a back edge of 7 nanosec decay time.

(7.2) Time Calibrators

For measurements where the time-base accuracy is important, it is necessary to employ some method of time calibration which can be recorded photographically. For moderate speeds, calibration pips which are sufficiently narrow to be superimposed on the trace can be generated from crystal oscillators. When this is undesirable owing to the mark/space ratio of the calibration pips, a double-beam oscillograph tube may be used with a common time-base; or with a single-beam tube a calibration trace may be recorded either immediately before or after the event. Non-crystal-controlled oscillators of adequate stability can also be used, and these have the advantage that they may be switched on by the signal, thus producing stationary pips when repetitive test waveforms are being observed. Above 50 Mc/s it is very difficult to generate the calibration waveform as pip and sine waves are normally employed. Lecher bar oscillato are suitable in the range 100–1 000 Mc/s, and cover most ultra high-speed requirements. The stability of cable delays can b used to advantage, especially in recycling generators[33] in whic each calibration pip produced is regenerated after an interva set mainly by a cable delay.

(8) CONCLUSIONS

There have been substantial advances in the past 10 years i the design of sealed c.r.t.'s suitable for recording high-spee single-shot transients. Many research workers have investigate ways of increasing the writing speed of the c.r.t. without resulta loss in sensitivity. One method has been to use a high degre of post-deflection acceleration. Another method relies on th production of a very small spot size, which enables a hig sensitivity to be obtained in terms of volts per trace width. Th method has been combined with travelling-wave deflection t produce a system of wide bandwidth as well as high sensitivit A third method under investigation uses the phenomenon o field-enhanced luminescence from the phosphor of the c.r. This method relies on the application of a field across the pho phor, resulting in electron multiplication within the phosph and an increase in luminous output. In yet another meth under investigation the face of the c.r.t. consists of a large numb of glass fibres cemented together to act as light guides. Th object of this system is to dispense with the normal lens b making it possible to place the recording film in contact with th front face of the c.r.t. It is hoped that by these means th optical efficiency will be increased.

There is a great need for an increase in video-amplifier ban width. Although the use of travelling-wave amplifiers feasible, the complexity of the system and non-linearity like to be introduced by the necessary modulation and demodulati limit their use in video systems, i.e. where the bandwidth required to extend down to a low frequency. There does n appear to be any likelihood of a substantial increase in th bandwidth of distributed amplifiers. However, the advent high-speed travelling-wave c.r.t.'s requiring only a few volts f maximum deflection has obviated the need for high-level ampl fiers. These c.r.t.'s often enable amplifiers to be dispensed wit altogether, since many detectors are able to provide a sufficie linear output voltage to drive the c.r.t. directly.

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A HIGH-RESOLUTION MEASURING SYSTEM USING COARSE OPTICAL GRATING

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SUMMARY

The paper describes a new measuring system, using 100 line/in optical gratings, which contains no moving parts and has been incorporated in both continuous-path and co-ordinate-setting control systems.

By a method of cyclic switching of four photocells and subsequent filtering of the combined photocell signals, an a.c. waveform is produced which can be compared in phase with a reference waveform, to determine the magnitude and direction of relative displacement between two gratings, and to provide linear interpolation within the pitch of the gratings.

Details of the optical, mechanical and electronic design are given, and the effects of variables on the accuracy of the system are discussed.

Interpolation accuracy within $\pm 2 \times 10^{-4}$ in using 100 lines/in gratings has been achieved in practical systems.

LIST OF SYMBOLS

e = Measurement error as percentage of grating pitch.
f = Switching frequency, c/s.
= peripheral drum velocity in line-pitches/sec.
g = A function.
i, i_A, i_B, i_C, i_D = Instantaneous currents.
i_1, i_2, i_3, i_4, i_p = Instantaneous photocell currents.
I, I_1, I_2, I_3, I_4 = Direct components of photocell currents.
I_c = Collector current.
I_{co} = Common-base collector leakage current.
K = A constant.
m, m_1, m_2, m_3, m_4 = Depths of modulation.
n = A number.
s = Grating ruling pitch, in.
t = Time, sec.
v = Relative grating velocity, in/sec.
α = Common-base current gain.
β = Spatial phase angle between correct and actual position of photocells in moiré fringe pattern, rad.
ϕ_e = Phase error, rad.
θ = Spatial phase angle, rad.
θ_e = Value of θ for maximum phase error, rad.
λ = Moiré fringe wavelength, in.
ϕ = Temporal phase angle, rad.
$\Phi_1, \Phi_2, \Phi_3, \Phi_4$ = Switching waveforms.
χ = Skew angle, deg.
ψ = Phase change through filter, rad.
ω = Angular velocity, rad/sec.

(1) INTRODUCTION

A position-measuring system is an essential part of most position-control systems as it provides feedback information to which command information is referred.

Some of the requirements are as follows:

(*a*) The measuring system should maintain the required accura over a wide range of system parameters to ensure reliability, ease manufacture and setting up.

(*b*) It should contain no moving parts in itself or in the dri circuits which might wear or limit operating speeds, and should reasonably small.

(*c*) It should be applicable to both continuous-path a co-ordinate-setting control with simple associated circuits.

(*d*) Zero setting should be simple and easily carried out.

(*e*) It should be readily adaptable to either linear or circul measurements.

The paper describes a practical measuring system, usi coarse optical gratings, which is designed to meet the requirements.

(1.1) Gratings

Gratings are regular parallel rulings or markings on gla glass-backed resin or reflecting metal. About ten years ago process[1, 2, 3, 4, 5] was developed at the National Physic Laboratory which enabled gratings of high quality and gre accuracy to be made, and Guild[11] and others showed how th could be used as precise measuring scales for laboratory industrial linear or circular measurements.

The technique depends on the fact that, when light is sho either through a pair of similar glass transmission gratings through a glass grating and reflected back from a similar met reflecting grating, with their rulings approximately parallel, t amount of light transmitted or reflected along a given pa varies according to the relative positions of the gratings. T intensity of this light can be measured by a photocell with suitable optical system. In most such linear measuring syster the two gratings consist of a long measuring grating with lin ruled across its length and a short transmission grating of t same ruling pitch used as an index.

In the transmission system described the index grating skewed slightly with respect to the measuring grating, produci broad light and dark bands approximately at right angles to t rulings, as shown in Fig. 1. The spatial distribution of lig intensity at right angles to the bands is approximately sinusoid but with special types of index grating the pattern may almost a triangular wave. The bands are known as *moiré fring* and if the angle between the grating rulings is χ, and the ruli pitch is s, the distance λ between two adjacent bands is given

$$\lambda = \frac{s}{2 \sin \frac{\chi}{2}} \quad . \quad . \quad . \quad . \quad . \quad . \quad ($$

As the measuring grating moves in a direction normal to rulings, the fringes move also; a movement of one ruling pit causes the fringes to move one fringe wavelength, λ. Wh individual rulings cannot be monitored photo-electrically, t broad fringes can be detected easily.

Early grating measuring systems counted complete frin cycles only, and had to use gratings with a ruling pitch equal the measuring resolution required. A later improvement divid

Written contributions on papers published without being read at meetings are invited for consideration with a view to publication.

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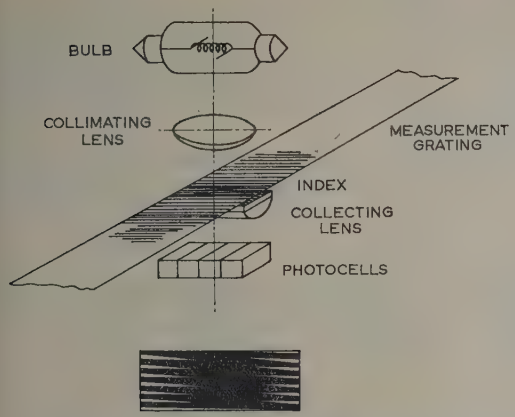

Fig. 1.—Diagrammatic sketch of optical system.
With enlargement of moiré fringe pattern formed by crossed gratings.

each fringe cycle into four parts, so that, to achieve a resolution of 1×10^{-4} in, fine gratings with 2 500 lines/in were required.

Fine gratings have been used successfully for a number of years[6] but suffer from many practical disadvantages. They are expensive and delicate, and require considerable skill in design to achieve good results. They also require precise mechanical and optical alignment, which may be difficult to maintain in some practical applications. At present it appears very difficult to use fine reflection gratings for measurement.

(1.2) Coarse Optical Gratings

Coarse optical gratings are here taken to be those in which diffraction effects in the formation of moiré fringes are negligible. In practical systems coarse gratings are usually ruled with less than 500 lines/in. The lines are usually opaque and equal in width to the spaces, and they can be easily reproduced photographically.

The 100 lines/in gratings used in the switched-sampling measuring system do not suffer from many of the disadvantages of fine gratings. They are cheap and the optical system is simple and easy to align. Coarse optical reflection gratings are robust. The mechanical and optical tolerances are many times greater than those of fine-grating systems, although their accuracy may be as high or higher. Shortly it is hoped to produce long lengths of 100 lines/in reflection gratings on stainless-steel tape, accurate to $\pm 1 \times 10^{-4}$ in/ft.

The disadvantage of coarse-grating measuring systems in the past has been that high-resolution measurement with them has entailed cumbersome interpolation methods. The switched-sampling measurement system described here has most of the electronic advantages of early coarse-grating systems without their mechanical disadvantages.

(2) DEVELOPMENT OF INTERPOLATION SYSTEMS

The use of 100 lines/in gratings in a high-resolution measuring system resolving to ± 1 or 2×10^{-4} in requires interpolation between fringe cycles to ± 1 or 2 parts in 100. One of the most accurate methods of interpolation is to measure the temporal phase difference between two square waves, and this is the method used in the measuring system here described.

Some time ago it was suggested[7, 9, 10] that, by introducing a continuously moving index grating, many of the disadvantages of the previous fixed-index system[6] might be overcome. In this case the continuously moving grating is used to set up two sets of moiré fringe patterns, one with the measuring grating and one with a fixed reference grating. Single photocells are used to detect the movement of both patterns, and a comparison of the phases of their outputs gives the required indication of the position and movement of the measuring grating.

The moving index grating may be in the form of a drum or disc.[9,10] If such a drum is driven at a line frequency f cycles per second and the measuring grating moves at v inches per second the output frequency associated with the moving moiré fringe pattern will be $f \pm v/s$ cycles per second, the sign depending on the direction of movement of the measuring grating.

It was later suggested[8] that similar effects could be achieved without a rotating grating by replacing a single light source by three lights positioned at intervals of $\lambda/3$, suitably positioned and excited from a 3-phase source, or by the inverse arrangement of replacing a single detector by three detectors and modulating the sensitivity of the detector by a 3-phase oscillator.

The interpolating measuring system here described is a development of the latter suggestion in that it avoids the complication of sine-wave modulation by using square-wave switching. It was first developed to be incorporated into a continuous control system in which command information is a square wave, phase modulated by digital means, the feedback is the output of the phase-modulated measuring system, and the position error is the phase difference between these two signals. As long life, low heat dissipation, small size, high speed and low cost are required, it was decided to switch a photocell system and not a lamp system.

It was also required to couple the measuring system to a computer or to a magnetic-tape link. With earlier systems[7,8,9,10] this would have necessitated an accurate phase locking of the effective peripheral drum velocity to a computer or tape reference frequency. Variation in tape speed, phase swinging in the motor, drum imperfections, etc., make this difficult.

Although the measuring system was first developed for continuous control it was soon realized that it is also particularly suitable for co-ordinate positioning and position-indicating systems.

(3) OPERATION OF THE MEASURING SYSTEM

The operation of the measuring system is first described with reference to the waveforms shown in Fig. 2 using skewed transmission gratings. The operation is similar if a reflection measuring grating is used.

Figs. 2(i)(*a*), 2(ii)(*a*) show a skewed index grating in combination with a measuring grating with zero relative phase shift of the grating lines referred to an arbitrary origin within any grating pitch of the measuring grating.

It is first assumed that there is no relative movement between the two gratings. The intensity of illumination in one cycle of the moiré fringe pattern is shown in Fig. 2(iii)(*a*). The depth of modulation is not 100% because of optical imperfections, skew, and the finite size of the photocells. The skew of the reference grating is adjusted to make one moiré fringe cycle a convenient length, e.g. $\frac{1}{2}$ in, and four photocells are positioned at $\lambda/4$ intervals across the cycle as shown by the small rectangles in Fig. 2(iii). Fig. 2(iv)(*a*) shows the magnitude of the currents which flow simultaneously in the four photocells. Each photocell is switched on in turn for half the switching period by a 4-phase square-wave switching waveform, Fig. 2(v)(*a*). Fig. 2(vi)(*a*) shows the magnitude and duration of the switched currents in the photocells. These four currents are now summed to form a stepped waveform, Fig. 2(vii)(*a*), which has a periodic time equal to that of the switching waveform, as it has been assumed that there is no relative movement between the gratings.

It should be particularly noted that, unlike the previous diagrams, which are spatial, Fig. 2(v) and the following ones are temporal, the time of one cycle being the 4-phase cyclic switching time, e.g. 500 microsec.

After a.c. coupling, the resulting symmetrical a.c. signal, which contains no d.c. component, is amplified and integrated to form the waveform shown in Fig. 2(viii)(*a*). One or more integrators are generally used instead of a filter which rejects all harmonics, because the phase shift through an integrator is independent frequency. This waveform is then squared in a circuit whi triggers accurately when the waveform crosses the mean lev The resulting square waveform is the measuring-system outp waveform, which is subsequently compared in phase with reference waveform.

The temporal phase difference between this square wavefo and a reference waveform, such as that used to switch photoc

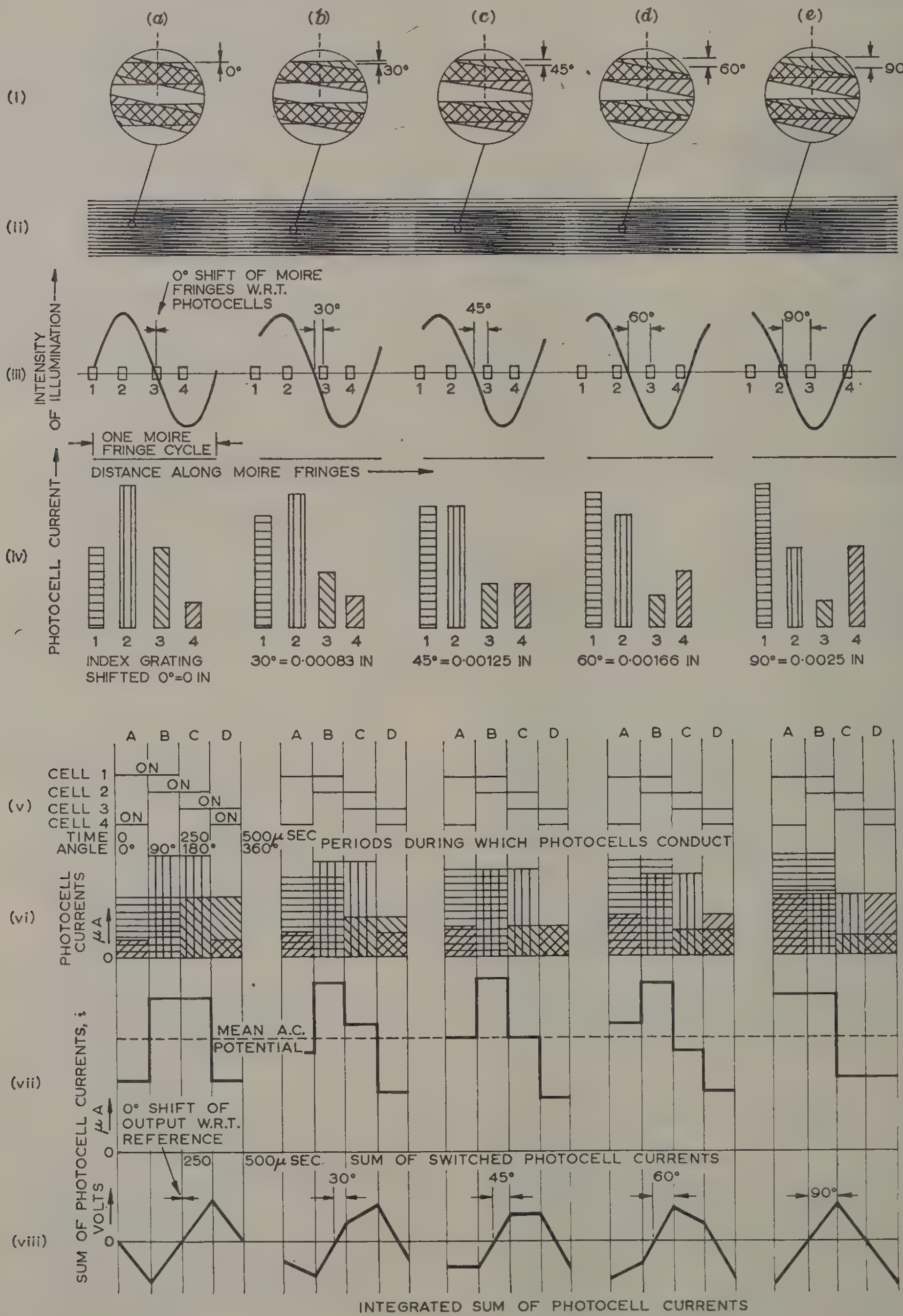

Fig. 2.—Switched-sampling measuring system waveforms.

No. 1, changes linearly with the spatial phase difference between the lines on the index grating and those on the measuring grating. In Fig. 2(*a*) this phase difference is zero.

Figs. 2(*b*)–(*e*) show equivalent waveforms for phase shifts of 30° to 90°, i.e. $8{\cdot}33 \times 10^{-4}$ in to $2{\cdot}5 \times 10^{-3}$ in for 100 line/in gratings.

It is shown in Section 5 that with certain practical shapes of wave pattern and types of filter, the temporal phase difference between these waveforms is an exact measure of the spatial phase difference between the lines on the two gratings. Theoretically an integrator should be used if the effective fringe pattern is triangular, and a filter should be used if the pattern is sinusoidal, but errors produced by using an integrator are in all cases small.

If the reading head containing the reference grating now moves over the measuring grating at a velocity $\pm v$ inches per second, the spatial moiré fringe pattern moves over the four photocells at a frequency v/s lines per second, and if the switching frequency of the photocells is f, the temporal frequency of the waveform Fig. 2(viii) is $f \pm v/s$ cycles per second.

(4) THE SWITCHING WAVEFORM

The switching waveform is usually a square wave of equal mark/space ratio although other ratios, for example 1 : 3, can be used. The advantages of square-wave switching compared with sine-wave modulation are as follows:

(*a*) The square wave can be accurately generated and locked to a higher frequency by digital subdivision.

(*b*) The switching waveshape requirements are relatively easy to achieve. The rise and fall time must be less than a fraction of 1% of the switching cycle, and the amplitude must be adequate for switching.

Since this is a sampling system, the frequency, f, of the switching waveform must necessarily be chosen so that the frequency $f - v/s$, at which relative position is indicated, is high enough to meet the requirements of the output system. Systems have been produced operating at switching frequencies in the range 250 c/s to 2·5 kc/s.

The practical upper limit to the frequency f is set by the switching time of switching transistors, which is less than 1 microsec.

(5) SIMPLE STATIC THEORY

We shall assume that the system is at rest.

The four photocell outputs are i_1, i_2, i_3 and i_4.

These are switched sequentially into a common summing circuit to give a resultant signal $i\,(\omega t)$, where ω is the angular frequency of the sampling cycle. This cycle may be divided into four equal periods, and the corresponding values of $i\,(\omega t)$ may be written i_A, i_B, i_C and i_D such that

$$\left.\begin{aligned} i_A &= i_1 + i_2 \\ i_B &= i_2 + i_3 \\ i_C &= i_3 + i_4 \\ i_D &= i_4 + i_1 \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

It will be noted that the pair of photocell outputs i_1 and i_3 are switched alternately into the output circuit and that these two signals are never present simultaneously. The same holds for i_2 and i_4. The alternating component of the resulting stepped waveform can therefore be resolved into two square-wave components of amplitudes proportional to $i_1 - i_3$ and $i_2 - i_4$, respectively, and displaced in phase by $\pi/2$ rad.

In filtering out the harmonics, the two components become two sine waves in quadrature, with amplitudes still proportional to $i_1 - i_3$ and $i_2 - i_4$, and the output becomes proportional to

$$i_f(\omega t) = (i_1 - i_3)\cos(\omega t + \psi) + (i_2 - i_4)\sin(\omega t + \psi) \quad . \quad (3)$$

where ψ is the phase change introduced by the filter. These two components are in fact never separated but form a resultant sine wave proportional to $\sin(\omega t + \psi + \phi)$, where

$$\phi = \arc\tan\frac{i_1 - i_3}{i_2 - i_4} \quad . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

If the spatial waveform is sinusoidal and the photocells have identical characteristics, we may write

$$\left.\begin{aligned} i_1 &= I(1 + m\sin\theta) \\ i_2 &= I\left[1 + m\sin\left(\theta + \frac{\pi}{2}\right)\right] = I(1 + m\cos\theta) \\ i_3 &= I[1 + m\sin(\theta + \pi)] = I(1 - m\sin\theta) \\ i_4 &= I\left[1 + m\sin\left(\theta + \frac{3\pi}{2}\right)\right] = I(1 - m\cos\theta) \end{aligned}\right\} \quad . \quad (5)$$

where I is the mean output level and m the depth of modulation.

Hence
$$\left.\begin{aligned} i_1 - i_3 &= 2mI\sin\theta \\ i_2 - i_4 &= 2mI\cos\theta \end{aligned}\right\} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

giving
$$\phi = \arc\tan\frac{i_1 - i_3}{i_2 - i_4} = \arc\tan\frac{2mI\sin\theta}{2mI\cos\theta} = \theta \quad . \quad (7)$$

The temporal phase angle, ϕ, of the filtered output is thus equal to the spatial phase angle, θ, and the output signal $i_f(\omega t)$ passes through zero whenever $\omega t + \psi = n\pi - \theta$.

If, instead of filtering $i(\omega t)$ to remove all the harmonics, we merely integrate, the two square-wave components of $i(\omega t)$ become two symmetrical triangular components in quadrature, again with their amplitudes proportional to $i_1 - i_3$ and $i_2 - i_4$.

The integrated output is thus proportional to

$$i_t(\omega t) = (i_1 - i_3)\,g(\omega t) + (i_2 - i_4)\,g\left(\omega t - \frac{\pi}{2}\right) \quad . \quad (8)$$

where $g(\omega t)$ is the symmetrical triangular function of time. If now the spatial waveform, instead of being sinusoidal, is also of a symmetrical triangular form, and the photocells are again assumed to have identical characteristics, we may write

$$\left.\begin{aligned} i_1 &= I[1 + mg(\theta)] & i_2 &= I\left[1 + mg\left(\theta + \frac{\pi}{2}\right)\right] \\ i_3 &= I[1 + mg(\theta + \pi)] & i_4 &= I\left[1 + mg\left(\theta + \frac{3\pi}{2}\right)\right] \end{aligned}\right\} \quad . \quad (9)$$

since the same function, $g(\theta)$, describes the spatial triangular wave. From the symmetry of $g(\theta)$ we have

$$\left.\begin{aligned} g(\theta) &= -\,g(\theta + \pi) = -\,g(-\theta) \\ g(\theta) &= g(\pi - \theta) \\ g\left(\frac{\pi}{2} + \theta\right) &= g\left(\frac{\pi}{2} - \theta\right) \end{aligned}\right\} \quad . \quad . \quad (10)$$

Hence we may write

$$i_t(\omega t) = 2mIg(\theta)g(\omega t) + 2mIg\left(\theta + \frac{\pi}{2}\right)g\left(\omega t - \frac{\pi}{2}\right) \quad . \quad (11)$$

It will be seen that this passes through zero whenever

$$\omega t = \frac{(2n + 1)\pi}{2} - \theta \quad . \quad . \quad . \quad . \quad (12)$$

Inspection of the geometry of the $i_t(\omega t)$ waveform shows that, although the shape changes as θ changes, there are always only two instances per cycle at which $i_t(\omega t) = 0$.

As before, the phase of the sampling cycle, ωt, at which the output $i_t(\omega t)$ passes through zero in a given direction is therefore uniquely and linearly dependent on the spatial phase angle, θ.

It should be emphasized that the above analysis applies only when the system is at rest. If one grating moves with respect to the other, provided that $f > v/s$, then the mean frequency of the output changes linearly with the rate of change of θ. The dynamic performance of the measuring system is mentioned briefly in the Appendix.

(6) THE OPTICAL SYSTEM

In this paper a transmission version of the optical system is described using skewed 100 lines/in transmission gratings. Reflection optical systems are in process of development; it is considered that they will ultimately replace systems using coarse glass gratings.

Fig. 1 shows a 6-volt 15-watt festoon lamp run at about 5 watts, a 40-diopter collimating lens, an index grating, a cylindrical collecting lens and four photocells mounted in the reading head. The reading head is compact and robust, the overall dimensions being 3 in × 2¼ in × 2¼ in. It is mounted on one of two sliding members, for example on the saddle of a machine-tool. The long grating is mounted in a special holder on the other member, for example on the table. The gap between the emulsion sides of the gratings is set to about 3×10^{-2} in, and the gratings used are 18 or 24 in long. They can be mounted in special holders which may contain up to three gratings end to end, forming a scale up to 6 ft long. Small gaps are left between the ends of gratings to allow for thermal expansion. Provided that adjacent gratings are accurately aligned with respect to each other, a gap involving the omission of one grating line introduces negligible error.

The photocell sensitivities can be balanced by screws of 2×10^{-2} in diameter which can be set to obscure part of the photocell surface.

The index grating, the collecting lens, the sensitivity screws and the photocells are mounted in a cylindrical holder in the reading head. The skew of this holder can be accurately set by a long removable tommy bar, so as to match the spatial phase of the photocells to the moiré fringe pattern.

(7) THE PHOTOCELLS

The system obtains its positional information from the relative amplitude of the signals from four photocells, and to give this information reliably they must fulfil certain conditions:

(*a*) The cells must be sensitive to tungsten-filament light.
(*b*) The light-intensity/photo-current characteristic must be linear.
(*c*) The four cells should have reasonably well-matched sensitivities.
(*d*) They must be thermally and time stable.

Silicon photo-voltaic diodes fulfil these requirements reasonably well. The cells are linear current generators if the load resistance is low enough; the sensitivity varies about ¼% per deg C; the time stability is expected to be good; and quartets of cells can be matched to within 10% of a mean sensitivity.

The cells are operated as current generators at about a 30 μA mean level. The upper limit of the load resistance is about 5 kilohms if linearity is to be preserved under these conditions.

(8) CIRCUITS

The circuits developed for the measuring system are co[mpletely] transistorized. They consist of the following parts.

(8.1) Switching Waveform Generator

A 4-phase switching square waveform of f cycles per seco[nd] with fast transitions, accurate phasing and unit mark/space ra[tio] is made by dividing a frequency of $4f$ by 4 using conventio[nal] transistor circuits. As shown in the block diagram, Fig. 3,

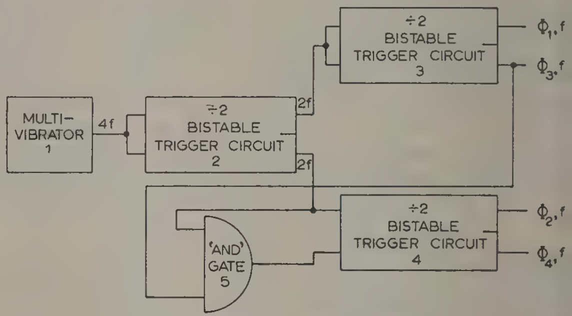

Fig. 3.—Switched waveform generator.

multivibrator, 1, produces a waveform of, for example, 8 kc[/s]. The shape of this waveform is unimportant so long as the positi[ve] transitions are adequate to drive a divide-by-2 stage, the outpu[ts] of which are anti-phase square waves of unit mark/space ra[tio] of 4 kc/s. These two outputs feed two more divide-by-2 stag[es] 3 and 4, via an 'and' gate, 5, so that the outputs Φ_1, Φ_2, Φ_3, [Φ_4] consist of a 4-phase square-wave system of constant pha[se] rotation. Gate 5 is necessary if the direction of phase rotati[on] is not to depend on the initial setting of the flip-flops when t[he] circuit is switched on.

(8.2) Silicon Photo-Voltaic Diode Switching Circuit

To obtain satisfactory operation from silicon photo-volta[ic] cells, the circuit design must fulfil two conditions. First, [in] order to obtain a linear relationship between photo-current a[nd] intensity of illumination, it is necessary to use a low-impedan[ce] load. A suitable load is the input impedance of a germaniu[m] transistor used in the common-base configuration. The seco[nd] requirement is that the current through the photocell must n[ot] be interrupted when the current is sampled by the switchi[ng] waveform. This is because, if light shines on an open-circuit[ed] silicon photo-voltaic diode, a charge is built up across the p[-n] junction, and when the cell is short-circuited the dissipation [of] the charge forms a pulse of output current.

These two requirements are met in the switching circuit sho[wn] in Fig. 4. Each photocell (e.g. D_5) is connected between t[he] base and emitter of a common-base transistor (e.g. T_1).

The collector current I_c of this transistor is given by

$$I_c = \alpha i_p + I_{co} \quad . \quad . \quad . \quad . \quad . \quad (1$$

Typical Characteristics of MS1 Silicon Solar Cells

Dimensions	Peak spectral response, μ	Output characteristics at 100 ft-candles		
		Open-circuit voltage	Short-circuit current	Rise-time
0·09 in × 0·19 in	approx. 0·8	400 mV	22·5 μA	about 1 microsec

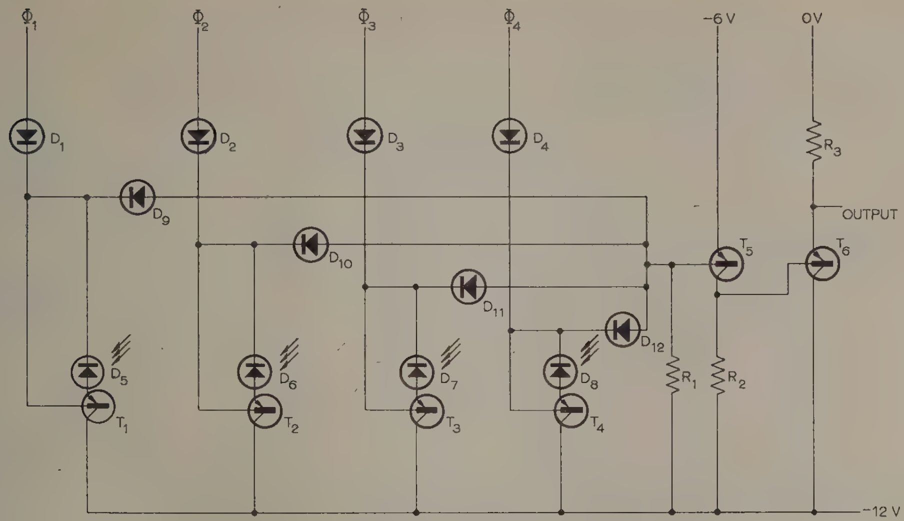

Fig. 4.—Photocell switching circuit.

D_1–D_4	Silicon diodes, ZW2	T_1–T_4	Transistors, OC44
D_5–D_8	Silicon photo-diodes, MS1	T_5	Transistor, OC201
D_9–D_{12}	Germanium junction diodes, OA5	T_6	Transistor, OC71
R_1	150 kΩ	Φ_1–Φ_4	Four-phase switching inputs
R_2, R_3	1 kΩ		

A high-gain r.f. transistor is used, in which I_{co} is small and α is very nearly equal to unity and is nearly independent of temperature. The collector current is therefore very nearly equal to the photocell current.

The collector current is switched between two circuits by means of a diode gate (e.g. D_1 and D_9). The switching waveforms are taken to the negative ends of the diodes D_1–D_4 and go positively and negatively with respect to the base of T_5.

When a switching waveform, e.g. Φ_1, is negative, D_1 is cut off and current flows from the base of T_5, through D_9, into the emitter of T_1, the magnitude of the current being proportional to the intensity of illumination on D_5. When the switching waveform is positive, D_1 conducts and D_9 is cut off. Thus the action of switching waveform Φ_1 is to switch a current having the same magnitude as the photocell current out of the base of an amplifying transistor during the negative parts of the switching cycle.

The action of the switching waveforms Φ_2, Φ_3, Φ_4 is similar. The switched currents representing each photocell current are summed at the base of T_5, and the stepped waveform of Fig. 2(vii) is produced across the collector load R_2. An emitter-follower, T_6, is used as an output stage.

Using a 15-watt lamp run at 5 watts the individual photocell currents are approximately 30 μA, and the output of the circuit is approximately 1 volt (p–p).

The switching circuits are located in the reading head, they are electrostatically screened and the supplies are filtered. Some integration of the stepped waveform by cable capacitance is unimportant as this waveform is integrated before use.

(8.3) Integrating and Squaring Circuits

In the circuits described (Fig. 5), an integrating amplifier is used. The amplifier consists of a common-collector input stage T_1, which is a.c. coupled to a common-emitter amplifier, T_2. The input signal is fed into the amplifier via R_1 and shunt feedback is applied by a capacitor, C_1. The time-constant of the integrator produced by the capacitive feedback, C_1R_1, is adjusted so that the output is approximately 6 volts (p–p) when the input is 1 volt (p–p).

The output after integration of the stepped waveform is shown in Fig. 2(viii). This output is taken to a squaring stage which triggers accurately at the mean alternating potential of the input waveform, so that changes in amplitude of the input waveform do not appreciably affect the trigger point and hence introduce a temporal phase error.

The circuit operates in the following way:

The integrator output is capacitance coupled to an emitter-follower, T_3, which is d.c. coupled to a squaring stage, T_4. The values of R_8 and R_9 are selected so that the d.c. output of T_3 biases T_4 to its linear operating region.

In order to obtain fast transitions, the output is d.c. coupled into an inverting stage, T_5, and the output of this is a.c. coupled to the base of T_4. The resultant positive feedback produces a fast leading edge on the square-wave output which is suitable for the triggering of counting circuits.

(8.4) Measurement of Temporal Phase Shift between Output Square Wave and a Reference Square Wave

Different methods of temporal phase comparison are used depending upon whether the measuring system is required to indicate position digitally or by analogue means or to feed a co-ordinate positioning system, profiling system or phase-to-pulse conversion system.

The phase comparator described consists of a simple flip-flop with two inputs into which are fed a reference signal (which could be, for example, phase 1 of the switching waveform) and the output of the squaring circuit. The flip-flop is turned on by the rising edge of the reference waveform and off by the rising edge of the feedback waveform. If these waveforms are displaced by 180° the d.c. component of the flip-flop output waveform is taken as zero.

As the temporal phase shift of the output waveform from the squaring circuit is advanced or retarded with respect to the reference waveform, i.e. as the position of the reference grating

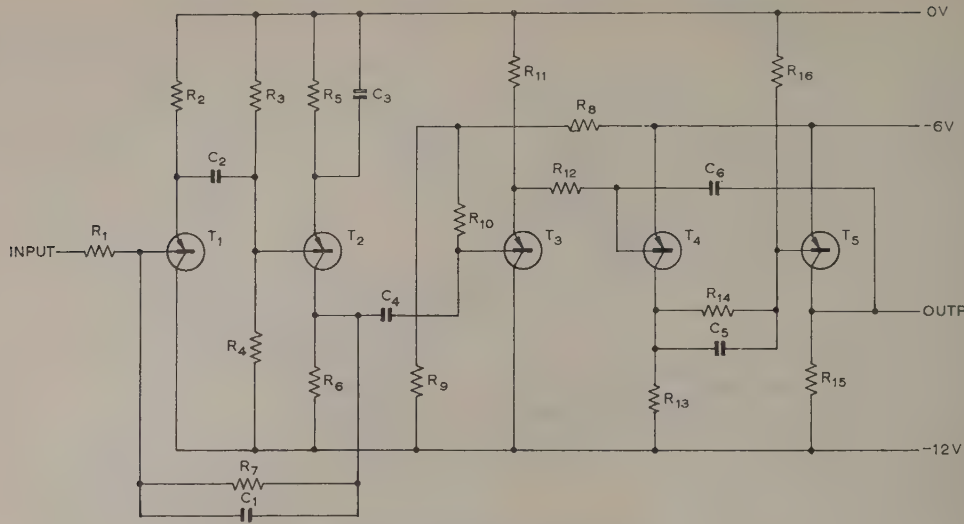

Fig. 5.—Integrating and squaring circuit.

R_1, R_2	10 kΩ	R_9	5·6 kΩ	C_1	0·001 μF
R_3	1·2 kΩ	R_{10}	10 kΩ	C_2	0·5 μF
R_4	12 kΩ	R_{11}	2·2 kΩ	C_3	8 μF
R_5, R_8	820 Ω	R_{13}, R_{15}	1·5 kΩ	C_4	0·1 μF
R_6, R_{12}	4·7 kΩ	R_{14}	18 kΩ	C_5, C_6	0·002 μF
R_7	470 kΩ	R_{16}	68 kΩ	T_1–T_5	Transistors, OC76

is moved forwards or backwards with respect to the measuring grating, the d.c. component of the flip-flop waveform becomes proportionally positive or negative respectively.

(8.5) Zero-Setting Circuit

The temporal phase shift between the reference waveform and the squaring-circuit waveform can be set to zero, for any relative spatial displacement of grating lines, by shifting the phase of the reference waveform.

This can be done by a 360° voltage divider tapped at 0°, 90°, 180° and 270°. These points are fed by the switching waveforms Φ_1, Φ_2, Φ_3 and Φ_4 respectively, and as the slider is rotated, the slider output waveform is similar to the stepped waveform, and changes in the same manner. This stepped waveform is converted into a square wave of temporal phase which varies with the angular position of the slider, using integrating and squaring circuits as described in Section 8.3. One revolution of the slider is equivalent to 0·01 in displacement when using 100 lines/in gratings.

The circuits described have been designed to operate satisfactorily up to 35° C.

(9) FACTORS AFFECTING ACCURACY OF MEASURING SYSTEM

At first sight it might be thought that considerable practical difficulty must be experienced in aligning photocell sensitivities, light-modulation depths, skew, etc., in order to achieve adequate overall accuracy. Reference to the Appendix will show, however, that under static conditions the tolerances on most of these variables are wide and that, with normal engineering tolerances, it should be quite possible to keep the contribution of individual sources of error to less than 1% of the grating pitch.

It will also be seen that the effects of several potential sources of error are amenable to mutual cancellation, if a suitable setting-up technique is adopted. As indicated in the next Section, practical results bear this out and show that overall interpolation accuracies within ±2% can be achieved consistently.

(10) EXPERIMENTAL RESULTS

(10.1) Method of Setting up the Reading Head

The following method of setting up balances the mean outputs of the photocells and sets the skew.

(*a*) With the head removed from the measuring grating, the four photocell currents are made equal by screws adjusting the effective photocell areas.

(*b*) The reading head is mounted and the skew of the reference grating is set by obtaining the best compromise balance between the outputs of the two out-of-phase pairs of cells.

Fig. 6 shows a graph of the temporal phase error between the output waveform and the reference wave plotted against actual displacement, at 21·2° C. The maximum interpolation error over typical grating cycles did not exceed $\pm 1{\cdot}0 \times 10^{-4}$ in. The progressive error was not greater than $\pm 2{\cdot}5 \times 10^{-4}$ in and the total error was not greater than $\pm 3{\cdot}5 \times 10^{-4}$ in over a 54 in length.

Actual relative displacement was measured by a calibrated line standard and micrometer microscope to an accuracy of $\pm 5 \times 10^{-5}$ in. Indicated relative displacement was measured by displaying the output square waveform on an oscillograph. This square wave was intensity modulated by a waveform of $20f$, from which the cell switching frequency, f, was derived. As the relative displacement of the two gratings was changed the output square wave was moved with respect to the bright spots until a bright spot coincided with the rising edge of the square waveform. A microscope reading of the line standard was taken at this setting. The rising edge was then moved to the next bright spot and the reading repeated. In this way indicated displacement can be read to better than $\pm 10^{-5}$ in.

(10.2) Control Systems Incorporating the Measuring System

Position-indicating systems or co-ordinate-setting systems giving meter indication within one grating cycle can be produced by indicating the d.c. component of the phase comparison flip-flop. Complete grating cycles can be indicated on number tubes. Digital interpolation to 10^{-4} in can be achieved by causing the

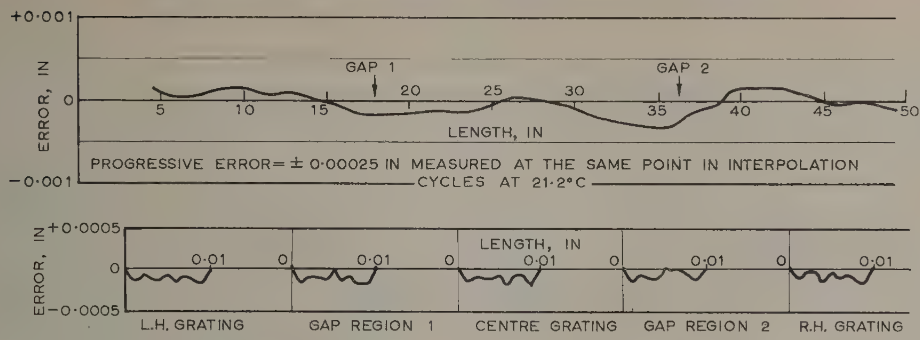

Fig. 6.—Interpolation error measured over typical interpolation cycles at 21·2° C.
Progressive and interpolation error for typical 54 in glass grating composed of three 18 in lengths with 0·015 in butt gaps at 21·2° C.
Total error, ±0·000 35 in for 54 in length.

ıase-comparison flip-flop waveform to gate a pulse train of)0*f* cycles per second. At intervals these pulses are fed into a ›unter which is cleared at corresponding intervals at the ends switching cycles. The number to which the counter counts equal to the relative displacement between the grating in eps of ten-thousandths of an inch.

The d.c. component of the flip-flop signal can also be used a servo error signal for continuous control systems. The ›mmand signal is obtained by phase modulation of either the ference signal or the waveform from which the switching signals e obtained. When the switching signals are modulated, the ference phase and the system output phase remain constant. his minimizes the frequency/phase-shift effect, but a common gnal generator cannot be used for reading-head switching in ore than one axis of control.

For some other applications such as phase-to-pulse con-ersion it is convenient to phase-modulate both reference and vitching signals simultaneously by digital means.

(11) CONCLUSIONS

The switched-sampling measuring system, in its present state f development, permits reliable interpolation to ±2 parts in)0. Thus coarse gratings can be used in co-ordinate-positioning ıd continuous control systems with considerable circuit conomy, while preserving adequately wide electrical and echanical tolerances. It uses no moving parts and is simple make both for angular and linear measurements.

Electronically it is a very flexible system because both the ference waveform and the switching waveforms can be phase-ıodulated by digital or analogue means. The system imposes serious limits on operating speeds for present machine-tool pplications.

The measuring system has been applied satisfactorily to osition indication, co-ordinate setting, contour milling and ore-measurement machine-tool control systems.

Although the measuring system has been described using)0 lines/in gratings, it could also be used with the same circuits ıd finer diffraction gratings, e.g. 1 000 lines/in, using optical chniques which are well established, although less simple, to roduce the moiré fringe pattern. In this way accurate inter-olation could probably be made to ±20 microinches, although ıe optical and mechanical tolerances would be smaller.

(12) ACKNOWLEDGMENTS

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(14) APPENDIX

In the simple static theory given above it was assumed that:

(i) The four photocells are perfectly matched as to sensitivity and mean level of illumination.
(ii) The four photocells are subject to the same degree of optical modulation.
(iii) The four photocells are correctly positioned with respect to the optical wave pattern.
(iv) The optical wave pattern is either truly sinusoidal, in which case a band-pass filter is used; or it is of a symmetrical triangular form, in which case an integrator is used.
(v) Measurement is made with the system at rest.

In practice there must necessarily be departures from these ideal conditions, and Tables 1 and 2 are given as some guide to the magnitudes of errors to be expected from various types of maladjustment.

The results given in Table 1 are mainly general and apply to departures from an ideal case in which a sinusoidal wave pattern is used with a band-pass filter.

The values given in Table 2 refer more specifically to a particular measuring system using 100 lines/in glass gratings with a nominal gap of 0·03 in between reference and measuring gratings.

Table 1

ERRORS CAUSED BY DEVIATIONS FROM IDEAL CONDITIONS

Conditions causing error	Phase-error equations	Magnitude of maladjustment	Resulting maximum percentage error, e
1. One photocell with abnormal sensitivity or illumination. $I_1 = KI \quad I_2 = I_3 = I_4 = I$	$\phi_e = \arc tan \frac{(K-1) + (K+1)m \sin\theta_e}{2m\cos\theta_e} - \theta_e$	$K = 1{\cdot}1,\ m = 1$	$\pm 1{\cdot}0$
	$m^2(K+3)\sin^2\theta_e + 2Km\sin\theta_e - (2m^2 - K + 1) = 0$	$K = 1{\cdot}1,\ m = 0{\cdot}5$	$\pm 1{\cdot}68$
2. Mismatch between inner and outer photocell pairs. $I_1 = I_4 = KI \quad I_2 = I_3 = I$	$\phi_e = \text{arc tan}\,\frac{(K-1) + m(K+1)\sin\theta_e}{(1-K) + m(K+1)\cos\theta_e} - \theta_e$	$K = 1{\cdot}1,\ m = 1$	$\pm 1{\cdot}06$
	$\sin\left(\theta_e - \frac{\pi}{4}\right) = \frac{(K-1)\sqrt{2}}{(K+1)m}$	$K = 1{\cdot}1,\ m = 0{\cdot}5$	$\pm 2{\cdot}07$
3. Mismatch between 'balanced' (antiphase) photocell pairs. $I_1 - I_3 = KI \quad I_2 = I_4 = I$	$\phi_e = \text{arc tan}\,(K\tan\theta_e) - \theta_e$ [see Note (*a*)] $\theta_e = \text{arc tan}\sqrt{\frac{1}{K}}$	$K = 1{\cdot}1$	$\pm 0{\cdot}76$
4. Mismatch of significant leakage currents in photocell circuits $I_{co1} \neq I_{co3} \quad I_{co2} \neq I_{co4}$	$\phi_e = \text{arc sin}\,\eta$ where $\eta = \frac{\sqrt{[(I_{co1} - I_{ce2})^2 + (I_{co2} - I_{co4})^2]}}{2Im}$	$\eta = 0{\cdot}1$	$\pm 1{\cdot}6$
5. Simultaneous mismatch of optical modulation depths of 'balanced' photocell pairs. $m_1 \neq m_2 \neq m_3 \neq m_4$ $I_1 = I_3 = KI$ $I_2 = I_4 = I$	$\phi_e = \text{arc tan}\left[\frac{K(m_1+m_3)}{m_2+m_4}\tan\theta_e\right] - \theta_e$ [see Note (*b*)] $\theta_e = \text{arc tan}\sqrt{\frac{m_2+m_4}{K(m_1+m_3)}}$	$\frac{K(m_1+m_3)}{m_2+m_4} = 1{\cdot}1$	$\pm 0{\cdot}76$
6. Incorrect spatial phasing of photocells due to small skew-angle error between gratings. [See Note (*c*)]	$\phi_e \simeq \text{arc tan}\,\frac{\sin\left(\theta_e + \frac{\pi\lambda\delta\mathcal{X}}{2s}\right)}{\cos\left(\theta_e + \frac{\pi\lambda\delta\mathcal{X}}{s}\right)} - \theta_e$	$\delta\mathcal{X} = 0{\cdot}001\,6$ rad $= 5{\cdot}5'$ of arc and assuming $\lambda = 0{\cdot}5$ in $S = 0{\cdot}01$ in	$\pm 1{\cdot}0$
7. Use of integrator rather than band-pass filter when spatial wave pattern is sinusoidal.	e.g. For $0 < \theta < \frac{\pi}{4}$ $\phi_e = \frac{\pi}{2}\left(\frac{\sin\theta_e}{\sin\theta_e + \cos\theta_e}\right) - \theta_e$ $\sin 2\theta_e = \frac{\pi - 2}{2}$, etc.	—	$\pm 1{\cdot}13$
8. Spatial wave pattern triangular rather than sinusoidal when using band-pass filter.	e.g. For $0 < \theta < \frac{\pi}{2}$ $\phi_e = \text{arc tan}\,\frac{\theta_e}{\left(\frac{\pi}{2} - \theta_e\right)} - \theta_e$ $\theta_e = \frac{\pi}{4}\left[2n + 1 \pm \sqrt{\left(\frac{4}{\pi} - 1\right)}\right]$	—	$\pm 1{\cdot}13$
9. Spatial wave pattern trapezoidal with flat portions each occupying one-sixth of total wavelength [see Note (*d*)]	—	—	$\pm 0{\cdot}96$

Notes.—(*a*) It will be seen that if the antiphase pairs of photocells are individually balanced, errors are independent of the depth of optical modulation.
(*b*) The form of this equation shows that a considerable measure of error cancellation is practicable.
(*c*) Errors due to incorrect skew setting can become of minor importance if four separately phased sections of index grating are used. With unjoined measuring gratings, the four index grating sections may be aligned parallel to the axis of the long grating. If, however, joined gratings are used, a square array of index grating sections presents certain advantages.
(*d*) The effect of such a trapezoidal wave pattern could result, for example, from the use of a measuring grating with a 2 : 1 mark/space ratio, in conjunction with a set of four unskewed index gratings each having a mark/space ratio of unity.

Table 2

MISCELLANEOUS ERRORS IN TYPICAL SYSTEM USING 100 LINES/IN GLASS GRATINGS WITH 0·03 IN GAP

Conditions causing error	Magnitude of maladjustment	Resulting percentage error, e
1. Variation in gap between index and measuring gratings when incident light is not strictly normal	Angle of incidence = 0·57°, variation in gap ± 0·010 in (see Note)	±1
2. Movement of lamp filament perpendicular to optical axis	±0·005 in with collimator of 1 in focal length	±1·5
3. Variation in lamp brightness when trigger level in phase comparator does not accurately agree with mean level of filtered signal	Trigger level = 0·1 × peak value of filtered signal; variation in lamp brightness = ±30%	±0·75
4. Differential thermal expansion of glass and cast iron for 24 in glass scale rigidly supported at centre only	±10° C	±4 at ends

Note.—An angle of incidence within ±0·57° corresponds to a tolerance ±0·010 in in filament position, if the lamp is working with a collimating lens of 1 in focal length.

General Note on the Dynamic Performance of the Measuring System

Instantaneous position is indicated only once per cycle of the output frequency. The significance of measurements made under dynamic conditions is thus subject to the usual limitations associated with sampling systems.

When the measuring system is used in a servo loop, provided that the output frequency is high compared with the servo cut-off frequency, the ordinary sampling errors do not persist long enough to affect the overall accuracy.

However, at a constant velocity there is a phase error per cycle which changes both in sign and magnitude over a time t which is the l.c.m. of $1/f$ and $1/(f + v)$. The average of the position error over the time t is exactly zero.

No rigorous analysis of the magnitude of this error has been made, but an experimental check of $v = \pm 0{\cdot}2f$ showed that no dynamic position error exceeded ±1%, which was the experimental accuracy. Also, in a contouring system using 100 lines/in gratings, where v was in the range $\pm 0{\cdot}4f$, there was no indication of dynamic position errors comparable with the $\pm 1 \times 10^{-3}$ in mechanical error expected from the machine.

An examination of the theoretical output waveforms for $v = +f$ and $v = +0{\cdot}5f$, where the positive sign represents phase advance of the output waveform, showed that the phase error in any cycle did not exceed ±5%. At $v = -0{\cdot}5f$ the phase error was zero.

DISCUSSION ON 'SUBSCRIBER TRUNK DIALLING'*

Before the WESTERN CENTRE *at* BRISTOL *9th November, the* TEES-SIDE SUB-CENTRE *at* MIDDLESBROUGH *2nd December*, 1959, *and the* NORTHERN IRELAND CENTRE *at* BELFAST *12th April*, 1960.

Mr. D. H. Mason (*at Bristol*): I am interested in learning how many sources of unit time impulses are employed, namely one per exchange or one common system for the whole country.

If the former is the case, how are all these units synchronized to ensure that the change of time-buying rate (night rate) occurs simultaneously throughout the country? One's imagination boggles at the idea of thousands of time generators changing their rate at the same moment all over the country.

Lastly, and in view of the new method described for timing and charging telephone calls, will telephone conversations be automatically charged at different rates should these extend from one rate-charge period to another?

Mr. J. C. Dallow (*at Bristol*): How does the s.t.d. system cater for those whose faculties are diminished through age or affliction? It may be difficult or impossible for such persons to dial the long series of digits which the new system demands.

The three tariff maps show that, as time goes on, the number of charging zones have been reduced, until under subscriber trunk dialling they will be down to four. Would it not be better to scrap the lot with the introduction of the new system, or if that is too sweeping, to have only two, namely local and trunk?

Mr. H. W. Long (*at Bristol*): One major difficulty with subscriber trunk dialling which has arisen at Bristol is the faulty trunk circuit. Records taken at Bristol on selected subscribers' lines show a fault rate of the order of 4%. Most of these are classified as 'no tones' and are due to faulty links in the chain, the instances of wrong numbers being very small. This will always be a problem even when cable and amplifier failures have been cut to the irreducible minimum. An elaborate system of reporting failures of trunk circuits to the trunk test position has been developed and the routining of trunks immediately prior to the busy hour will enable most faulty circuits to be busied before they are seized by subscribers, but the real solution is to develop a type of trunk signalling which will give an automatic busying facility if the circuit is cut anywhere. There is one form of d.c. signalling which does this but its use is restricted to a.f. circuits, and a system suitable for circuits routed on high-frequency systems has not yet been developed. Such a system would enable the fault rate to be cut to under 1%.

Mr. R. W. Steel (*at Bristol*): In the electricity supply industry, meters have to be tested and certified as correct within the limits provided for in the Electricity Meters Act, 1936. Are there any similar statutory provisions for certifying the charging equipment used by the Post Office for calculating subscribers' accounts, and if so, what are the accuracy limits?

* BARRON, D. A.: Paper No. 2795 R, November, 1958 (see 106 B, p. 341).

Mr. F. P. Phillips (*at Bristol*): Will the equipment described be comparable with the initiative of a human operator in finding alternative routes when major damage to cable and equipment occurs?

Mr. N. Burley (*at Middlesbrough*): The author's hopes of 75% of the trunk calls being dialled by 1970 requires a programme for the installation of s.t.d. equipment which will call for very large capital expenditure.

As long ago as 1912, after the successful conversion of the Epsom Exchange to automatic working, the Administration decided that the future policy of the Department was the conversion of the manual system to automatic exchanges, and yet to-day, almost half a century afterwards, there are still more manual telephones in the United Kingdom than there were in 1912. This hardly gives one assurance that the conversion to s.t.d. working can be carried out during the next ten years unless Treasury restrictions on capital expenditure are greatly relaxed, for it is these which have so severely retarded progress.

The programme could never have been contemplated but for the building up of the trunk transmission system during the last 25 years. This has enabled many hundreds of trunk circuits to be provided either by means of coaxial cable or on a v.h.f. network by the process of frequency translation. So successful has been the skill of the designers of transmission equipment that the provision of long-distance circuits is now cheaper than it was 30 years ago on a per-circuit basis. The progress made by foreign administrations in converting their systems to s.t.d. working makes it essential for the Department to pursue energetically a similar policy if this country is not to lose its present position as having an advanced system of telecommunications.

In so far as the Middlesbrough telephone area is concerned, the programme now being commenced should result in the subscribers on Middlesbrough Main Exchange and Redcar having s.t.d. facilities by December, 1960. This will be followed, within the following 12 months, by Norton and Stockton subscribers. Many other exchanges throughout the area are then scheduled to be provided with the equipment during the following three to four years, but past experience of policy reversals makes one apprehensive of specifying too precisely the programme.

It is intended that register-translators shall be provided only at the Middlesbrough Main Exchange and that these will be used by the subscribers at the other exchanges, where the only equipment required to be provided will consist of timing mechanisms for operating subscribers' meters. It is proposed to convert the dialling code 'O' for obtaining the operator to the code '100' in Middlesbrough and Redcar by May, 1960. It will also be necessary to convert the codes '7XX' for the numerous exchanges at present dialled by Middlesbrough subscribers to '9XX' for exchanges in adjacent charging groups.

Mr. E. Shaw (*at Middlesbrough*): The type of s.t.d. system which could be introduced by the Post Office had to operate the existing trunk network. This was built up for manual operation and designed to give a minimum number of switched points in a circuit by connecting each zone centre to all other zone centres. Is this the best network for subscriber trunk dialling, or does automatic control of the trunk network mean the loss of some of its best features, i.e. flexibility of routing and minimizing of breakdown effects, without exploiting the traditional features of automatic working, i.e. high speed, reliable switching, etc.? How will the trunk network now develop? Will more trunk switching be introduced and routing be rearranged to meet subscriber trunk dialling?

Mr. N. C. C. de Jong (*at Belfast*): Northern Ireland is well placed to develop subscriber trunk dialling, since 92% of its telephones are connected to automatic systems compared with some 80% elsewhere in the United Kingdom. A completely automatic system is planned by 1965. Subscriber trunk diallin will start at the Central Exchange for some 11 000 subscribers b the end of 1961, and within a year or so it will be extended t the whole of the Belfast district and several important place outside. By 1963 all the main towns will be served, over 60% of all lines will have s.t.d. facilities and some three-quarters o all trunk calls will be self-dialled.

The national codes for Northern Ireland are based on sig nificant letters in the names of the 12 switching centres whic will serve the 34 charging groups, e.g. OLO4 Londonderry OPN2 Portadown; OBE2 Belfast.

The personal touch in the friendly telephone concept will b promoted by retaining nine manual trunk boards as assistanc centres, but the economics of the scheme will be achieved b reducing the operating force from 700 to about 500; this i being carefully planned in harmony with the Trade Unions t minimize hardship to the permanent staff, none of whom wi be dismissed.

After 1963 the further extension here of subscriber trun dialling depends on the policy to be adopted for rural area Once this is settled the facilities could be made available through out Northern Ireland within a few years, so that there is prospect of a 100% s.t.d. system by 1967. The real difficulty i that of cost, and certain possibilities of minimizing it ar suggested:

(i) Employing a cheaper form of register-translator which woul provide for automatic connection only between centres having community of interest.

(ii) Providing facilities for meter pulses to be sent back to th local exchange over two junctions in tandem, thus avoiding th necessity for every exchange to have a direct route to its grou switching centre.

Next to cheaper calls the main advantage of subscriber trun dialling lies in speed, and no doubt future plans, such as for fast transit network, will provide for even faster operation. Th dialling of 11 or more digits is a comparatively slow process, an techniques are already available which are much faster. Th question is whether these will have to wait for electroni exchanges.

The popularity of subscriber trunk dialling is obvious fro the Bristol results, and extracts of interest are as follows:

90% said they preferred subscriber trunk dialling.

88% said they had no difficulty in distinguishing between th various tones.

88% said it was easy to find the required national codes in th booklet provided.

83% thought that the inclusion of letters in the national codes wa helpful.

75% said that the layout of the new accounts was easy to unde stand.

This development is a major step forward in promoting greate use of the telephone. At Bristol on s.t.d. routes traffic increase by 40%; even on non-s.t.d. routes there was an increase of 25% This, and other factors, demands a substantially increased num ber of long-distance circuits. It is as well, therefore, tha engineers have made these cheaper to provide, since, in thi respect, human operators are more economical and achieve greater circuit occupancy. It is suggested that this is a facto to be explored. Even more important is the greater efficienc of human operators in conditions of breakdown or congestion presumably automatic alternative route switching will be inco porated in s.t.d. equipment. The current emphasis on keepin personal contacts with the public as close as possible has le to a greater number of assistance centres than can be econom cally justified. There will therefore always be compelling pre sures to reduce the numbers, and experience elsewhere may b

most valuable. This greater dependence on automation makes it not only essential that the new equipment should be even more reliable but that existing switching equipment should be maintained at a higher standard.

Mr. J. W. Dorrington also contributed to the discussion at Bristol.

Mr. D. A. Barron (*in reply*): The synchronization of the time of change from the day to the night tariff raised by Mr. Mason will be carried out by a clock located at each centre equipped with register-translators—about 500 centres for the whole country. Daily checks will be made to ensure that the change has been carried out correctly. A call in progress when a tariff change is made will be charged for according to the tariff appropriate to each period of the call. For example, a call made when the day tariff is in force and extending until the night rate applies will be charged first at the day rate and subsequently at the night rate.

In reply to Mr. Dallow I should mention that subscriber trunk dialling is an additional service available to subscribers, and it will still be possible to pass calls through the operator for those who wish to do so. As in such cases the Post Office would not realize the economies which arise from subscriber trunk dialling, the cheaper s.t.d. tariff would not apply.

On the point raised by Mr. Steel, the meters used for recording the charge units for each subscriber are tested at regular intervals, while the accuracy of the pulse rates is checked at least twice each day.

The difficulty referred to by Mr. Long which arises when using the older designs of signalling systems is fully appreciated, and later designs of signalling systems now being installed provide an automatic busying facility. This facility is also provided on the a.f. circuits using loop-disconnect signalling between an originating exchange and its s.t.d. equipment at the group switching centre.

The provision of sufficient capital to carry out the full programme of conversion to automatic working mentioned by Mr. Burley of course involves other than purely technical considerations. However, a programme for full automatic working has been prepared, and the return for capital invested (particularly in subscriber trunk dialling) is sufficiently attractive to ensure a high priority for this project.

Several speakers, including Messrs. Shaw and de Jong, refer to the trunk network and its adaptation to meet the needs of a fully automatic system. The initial installations of s.t.d. equipment were designed to fit into the present trunk network, which, of course, represents a large investment in line plant and exchange equipment. It is appreciated, however, that with a fully automatic system a number of improvements can be incorporated, particularly in the speed and accuracy of setting up trunk calls. Equipment incorporating high-speed register-to-register signalling and fast switching techniques is at present being developed and will be brought into service as soon as possible. The new equipment will provide for automatic alternative routing, although some additional manual control may be necessary to meet the conditions envisaged by Mr. Phillips.

Mr. de Jong gives some interesting statistics of the extent to which subscriber trunk dialling will be provided in Northern Ireland during the next few years, and he rightly stresses the need for careful planning to minimize hardship to the staff. On the question of the extension of subscriber trunk dialling to rural areas the technical solutions have been worked out and the development of the necessary equipment is in hand. Although at first sight the ability to send back meter pulses over two junctions in tandem appears attractive, the technical problems involved in such a signalling system, together with the difficulty of obtaining, in all cases, a suitably low overall transmission loss for the two junctions, makes such a scheme uneconomic.

On the number and location of manual assistance centres, I have no doubt that experience during the next few years will be most valuable to us in framing our long-term policy.

PAPER AND MONOGRAPHS PUBLISHED INDIVIDUALLY

Summaries are given below of a paper and 12 monographs which have been published individually. The paper is free of charge; the price of the monographs is 2s. each (post free). Applications, quoting the serial numbers as well as the authors' names, and accompanied by a remittance where appropriate, should be addressed to the Secretary. For convenience, books of five vouchers, price 10s., can be supplied.

The Design of an Audio-Frequency Amplifier for High-Precision Voltage Measurement. Paper No. 3335 M.
S. Harkness and F. J. Wilkins, B.A., B.Sc.

The specification and design of an amplifier required for precision audio-frequency measurements is discussed. Formulae are presented for the 'ring of three' feedback circuit and a detailed analysis is made of appropriate feedback theorems. A multi-gain amplifier which extends the voltage range of the electrostatic voltmeters, used at the N.P.L. as basic a.c./d.c. transfer instruments, is described. An increase in this range of up to 1 000 times is provided and enables voltages between 60 mV and 60 volts in the frequency range 30 c/s–10 kc/s to be measured to a few parts in 10^4.

Unstable Electron Flow in a Diode. Monograph No. 402 E.
R. J. Lomax, M.A., Ph.D.

A perturbation analysis is used to demonstrate the instability of a type of electron flow in a plane diode known as C-overlap flow, which is predicted to be a possible flow in the approximation in which electrons are emitted from the cathode with a uniform velocity.

A New Approach to Kron's Method of Analysing Large Systems. Monograph No. 403 E.
R. Onodera.

Generally an electrical network is used as a model circuit for a physical field. This circuit is most complicated and the analysis of it is frequently difficult. 'Diakoptics', introduced by Kron, is very effective for analysis of the network. The method entails the operation of 'cutting', which is generally classified into open-circuiting and short-circuiting. The latter operation is the dual of the former, but, as far as the author is aware, has not yet been reported in any of the literature. A method is described, based on Kron's diakoptics, which uses the operation of open-circuiting, and is further extended to dual diakoptics treated by the operation of short-circuiting. Here a simplification of Kron's diakoptics is attempted and a dual method is introduced. It seems that this attempt goes backwards in tensor geometry but forward in combinatorial topology. The first intention of the paper is to show the duality between diakoptics and codiakoptics.

Audio Communication with Orthogonal Time Functions. Monograph No. 405 E.
Henning H. Harmuth, Dipl.Ing., Dr. Tech.

An audio signal may be decomposed into components having certain frequency and phase. Half the information of the signal is contained in the phase of the frequency components. Since the human ear is almost insensitive to phase, one may eliminate the phase information without causing a noticeable reduction in the signal quality. The elimination of the phase information implies a permissible reduction of the bandwidth required for the transmission of the signal to one-half and a reduction of the signal power by 3 dB

without increase of distortion due to additive white normal noise. A method for the elimination of the phase information is based on the decomposition of the audio signal by correlation with a set of orthogonal functions. These correlations yield for each orthogonal function one coefficient which is represented by a voltage. One may process these coefficients in analogue or digital computers to reduce further the bandwidth required for the transmission of audio signals with certain spectral distribution, e.g. voice signals.

Optimum Combination of Pulse Shape and Filter to produce a Signal Peak upon a Noise Background. Monograph No. 407 E.
H. S. Heaps, B.Sc., M.A.

The paper concerns the generation of a signal pulse for transmission through a propagating medium and its subsequent detection as a single peak after it is received upon a noise background. It is supposed that the propagating medium behaves as a linear filter and that the noise background is independent of the signal. The pulse is conveniently described as consisting of a central portion attached to a leading edge and a trailing edge. It is found that for a given length of central portion there is an optimum combination of transmitted pulse shape and predetection filter. The results are compared with those arising from the use of certain non-optimum systems, and it is found that the optimum system leads to a significantly high signal/noise ratio.

The results of the paper imply that for the range of parameters considered it is advantageous to transmit a succession of short pulses of a determined form rather than a single smooth pulse.

The Potential Distribution and Thermionic Current between Parallel Plane Emitters. Monograph No. 408 E.
F. H. Reynolds, B.Sc.(Eng.).

A thermionic system formed of two parallel plane emitters facing each other is analysed. By means of Tables, the distribution of potential between the emitters and the net space-charge-limited current which flows between them can be obtained for any values of emitter potential, temperature and saturation current density. The results are applied to practical problems including the thermo-electric and conductivity characteristics of oxide cathodes.

Propagation along Unbounded and Bounded Dielectric Rods: Part 1—Propagation along an Unbounded Dielectric Rod. Monograph No. 409 E.
P. J. B. Clarricoats, B.Sc.(Eng,), Ph.D.

The paper describes a method for evaluating the propagation coefficients of an unbounded lossless dielectric rod of infinite extent. The propagation coefficients are obtained as a function of the ratio of rod radius to free space wavelength, r_1/λ_0, for the three lowest modes of propagation possessing fields with θ dependence of the form $\varepsilon^{\pm j\theta}$. The method of solution enables the form of the complete mode spectrum to be identified. It is also established that the product of the rod permeability and permittivity $\mu\epsilon$, primarily determines the propagation behaviour; the ratio μ/ϵ has only a secondary effect. An expression is obtained for the distribution of transmitted power between the road and the surrounding space, and it is also demonstrated that the product $\mu\epsilon$ primarily determines this quantity. Correlation between power distribution and attenuation is demonstrated for a rod possessing small losses.

Propagation along Unbounded and Bounded Dielectric Rods: Part 2—Propagation along a Dielectric Rod contained in a Circular Waveguide. Monograph No. 410 E.
P. J. B. Clarricoats, B.Sc.(Eng.), Ph.D.

The paper describes both approximate and exact methods for evaluating the phase-change coefficients of a circular waveguide containing an axial dielectric rod. Close correlation is found between the phase-change coefficients of this waveguide structure and those of an unbounded rod over a wide range of rod radii. The correlation enables an unambiguous classification of the modes of an unbounded rod to be made. Exact and perturbation expressions are derived for total transmitted power, power distribution and attenuation in the dielectric-rod-waveguide structure. Correlation with the attenuation coefficients of an unbounded rod is again predicted. The application of the results of the paper to ferrite devices is briefly mentioned.

The Launching of Surface Waves by a Magnetic Line Source. Monograph No. 411 E.
C. M. Angulo and W. S. C. Chang.

The paper deals with the idealized problem of the excitation of surface waves along two infinite, identical, separate and parallel dielectric slabs by a magnetic line source (slot). The problem is reduced to a superposition of simpler ones, namely two parallel slabs with symmetrical excitation and two parallel slabs with antisymmetrical excitation.

The simplified problems are solved by the modal-analysis approach and the synthesis of the modal components is carried out to obtain the far fields excited by the magnetic line.

Radiation pattern, total power of the surface waves, total radiation loss and the efficiency of launching the surface wave are derived and computed numerically for various thicknesses of the slabs, various air-gaps between the slabs and different positions of the magnetic line source. The theoretical results indicate that, for each thickness of the slabs, high efficiency is obtained with an optimum location of the source and an optimum air-gap. Moreover, the thicker the slabs and the higher is the maximum efficiency. The thickness of the slabs and the air-gap are ultimately limited by the requirement that only one surface wave should exist for the structure.

The Output Spectral Density of a Detector operating on a F.M. C.W. Radar Signal in the Presence of Band-Limited White Noise. Monograph No. 412 E.
J. Lait, M.A., and A. J. Hymans, M.Sc.

A method of analysis suggested by Lawson and Uhlenbeck is used to examine the interaction between reference signal, echo and noise in the detector stage of a frequency-modulated c.w. radar receiver.

The paper gives a method of approximation which may be used in most practical cases and states the restrictions on system parameters for which the approximations are valid.

Expressions for spectral density are derived separately for both quadratic and linear detectors, and the question of optimum pre-detector bandwidth is examined. Finally, the authors consider how signal/noise ratio will be affected by choice of bandwidth in stages which follow the detector.

The Algebra and Topology of Electrical Networks. Monograph No. 414 E.
P. R. Bryant, M.A., M.Sc.

An integrating paper is presented, gathering together the various results and methods used in an elementary algebraic and topological study of electrical networks. Considerations are restricted to networks containing a finite number of lumped resistors, capacitors and self and mutual inductors. The result is a connected logical presentation leading, by use of the concepts of incidence, loop and cut-set matrices, to the corresponding nodal, loop and cut-set analysis methods.

The Indeterminacies of Measurements using Pulses of Coherent Electromagnetic Energy. Monograph No. 417 E.
R. Madden, B.S.

The measurements. on a single pulse function of electromagnetic radiation, of the position of a scatterer with polar co-ordinates R, ϕ are indeterminate in themselves. The positional indeterminacies are related by $\Delta R \Delta\phi \simeq \frac{1}{2}\lambda_t$, where λ_t is the wavelength of the source. The relation between the indeterminacies of range R and radial velocity V_r is found to be $\Delta R \Delta V_r \simeq \frac{1}{4}\lambda_t c$. It is shown that vector position and vector velocity are not measurable simultaneously, and it is suggested that the 3-dimensional measurement problem is basically limited to non-simultaneous measurements which have restrictions in the presence of multiple scatterers. Similarity is noted to the quantum-mechanical problem.

Frequency Spectrum Distortion of Random Signals in Non-Linear Feedback Systems. Monograph No. 419 M.
J. C. West, Ph.D., D.Sc., J. L. Douce, Ph.D., B.Sc., and B. G. Leary, B.E.

The paper considers the distortion introduced by torque saturation in a servo mechanism when responding to a randomly varying input signal, with emphasis on the spectral distribution of this distortion. It is shown that low-frequency errors are produced by the application of an input signal whose spectrum extends beyond the pass-band of the system.

An approximate expression is derived which enables the magnitude of the low-frequency errors to be evaluated in a simple manner. Experimental results are presented for a particular system.

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SCOTTISH CENTRE	*R. H. Dean, B.Sc.Tech.*
NORTH SCOTLAND SUB-CENTRE	*P. Philip*
SOUTH MIDLAND CENTRE	*H. M. Fricke*
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SOUTHERN CENTRE	*J. E. Brunnen*
WESTERN CENTRE (BRISTOL)	*A. H. McQueen*
WESTERN CENTRE (CARDIFF)	*E. W. S. Watt*
WEST WALES (SWANSEA) SUB-CENTRE	*O. J. Mayo*
SOUTH WESTERN SUB-CENTRE	*W. E. Johnson*

Members are asked to bring to the notice of the Court of Governors any deserving cases of which they may have knowledge.

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